1842 EEG/EETET TR 77-18

AFCS TECHNICAL REPORT

FM QUIETING CURVES

AND RELATED TOPICS

SYSTEM TECHNICAL APPLICATIONS FACILITIES
1842 ELECTRONICS ENGINEERING GROUP (AFCS)
RICHARDS-GEBAUR AIR FORCE BASE, MISSOURI

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This report describes the relation between readily measured frequency modulation radio transmitter and receiver parameters and the basic baseland signal and noise performance of the radio transmitter/receiver combination. Equations, graphs, and charts are provided to facilitate application.

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APPROVAL PAGE

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1. Introduction.

1.1 With all the articles, reports, and books written about frequency modulation, transmitters and receivers, why write another report? Despite the volumes written about frequency modulation equipment, several important features have been neglected. Other significant features have been described, but are spread over so many different sources or buried in so much extraneous material that it has been difficult to get the important information and put it in perspective. The principle interest in writing this report has been to give engineers and technicians the information necessary to understand and predict the basic performance of frequency modulation microwave radio terminals. A thorough knowledge of the performance of communication systems is required to adequately specify, develop, test, engineer, install, and verify proper operation of an item of communication equipment. For frequency modulation microwave radio terminals, this includes a knowledge of the baseband signal and noise properties of the terminal transmitter and receiver configuration. The noise at the baseband of an analog microwave receiver is a complex mixture of noise from such sources as radio terminal thermal and intermodulation noise, crosstalk, and frequency division multiplexer noise. This noise can generally be divided into two broad categories. The first category is noise which is independent of the signal applied to the terminal baseband. This noise includes the effect of thermal noise generated within the receiver, the phase noise of the transmitter, cable crosstalk from other baseband signals, radio frequency interference, and multiplexer tone leakage, power supply ripple, and miscellaneous sources. Since this noise has no relationship to the baseband signal, it can be measured with the baseband signal removed from the transmission equipment. It is called idle noise. The other main noise category includes the noise which is directly related to the frequency and power level of baseband signal. Although this noise includes such effects as baseband cable and radio terminal crosstalk, this type of noise is loosely termed intermodulation noise. This noise can only be measured with a signal applied to the transmission equipment baseband. The primary tool for predicting and evaluating idle radio noise is the frequency modulation (FM) slot noise quieting curve. The primary tool for evaluating radio intermodulation noise is the noise power ratio (NPR) bucket curve. Bucket curves can be used to predict and analyze such noise factors as waveguide and free space multipath effects and microwave terminal differential gain and phase. Utilization of bucket curves requires a full understanding of quieting curves. Unfortunately, present job duties preclude the investigation necessary to write a report on bucket curves. This report will only cover FM aspects of noise quieting curves and related topics. Nevertheless, quieting curves are quite important and deserve considerable attention. Idle noise sets the lower limits on the noise performance of an analog digital microwave radio terminal. The thermal noise component of a microwave receiver is the dominant factor in baseband signal and noise performance of the radio terminal for low power level radio signals.

Therefore, thermal noise of the terminal directly effects fade margin and radio link reliability as well as terminal noise performance. Although the present report is directed toward convencional frequency division multiplexed analog baseband transmission, the information is directly applicable with minor modification to the transmission of wideband video or digital quasi-analog signals over analog microwave equipment. An example will be given using quieting curve concepts to determine bit error rate performance of the three level partial response quasi-digital FM radio terminals currently being installed in the Digital European Backbone (DEB) digital wideband communications upgrade.

1.2 In the communications business, the object is to get information from one place to another with the least degradation in quality. Information that must be transmitted quickly over a long distance usually takes the form of a telephone call or a data circuit processed for transmission on a telephone circuit. These signals are routed over cables and controlled with various types of processing, monitoring, and routing. If the signals are transmitted very far, they almost invariably make their way to a microwave (M/W) transmission system for long distance bulk transmission. Prior to transmission, as many as several hundred separate telephone channels are combined (frequency division multiplexed or FDMed to produce a wide frequency range (wideband) baseband signal) for efficient processing. The M/W transmission system generally consists of several separate M/W links, each consisting of a radio path and two M/W radio terminals. At one terminal, the combined telephone circuits (baseband) are converted into a modulated radio signal with frequency as high as 11 gigahertz (GHz).

1.3 The modulated radio frequency (RF) signal passes through nonlinear transducers, imperfect transmission lines, and is transmitted through the air where it is susceptible to multipath, interference, and fading degradations. After transmission over several radio links, the baseband signal is reconverted (demultiplexed) into individual telephone circuits. While the telephone circuits exist as individual channels, they are processed by individual amplifiers and cables. They are susceptible to noise and other forms of distortion. The failure or degradation of a single telephone circuit, while undesirable, is not immediately hazardous to many customers. Failure or degradation of a single radio terminal, however, since it carries many telephone circuits, can have serious consequences. For this reason, considerable attention is given to M/W radio terminal performance.

1.4 As a radio terminal transmits a baseband signal from one location to another, the signal is easily influenced by nonideal characteristics of the overall transmission medium. The effect of these imperfections is to introduce noise into the reconstituted (demodulated) baseband at the last M/W terminal. The processing of baseband signals at a frequency modulation (FM) M/W radio terminal is complex. The noise degradation of the baseband signal can be grouped roughly into two

categories. The first type is noise which is primarily independent of the baseband signal itself. This noise can be introduced by M/W transmitter phase/frequency noise (phase jitter), various forms of interference and crosstalk, and internally generated ("front end") receiver thermal noise. For simplicity, these types of noise will be called simply "idle noise." The second type is noise which is directly related, among other factors, to characteristics of the baseband signal itself. Sources of this type of noise include waveguide echo and moding, multipath, differential gain and phase, amplitude modulation to phase modulation (AM to PM) conversion, as well as other forms of intermodulation distortion. For simplicity, this type of noise will simply be called intermodulation noise.

1.5 The primary method of characterizing the thermal noise performance of a M/W radio is through the use of what is called an FM slot noise quieting curve or simply quieting curve. The quieting curve is a plot of noise in the baseband of a M/W FM receiver as various levels of unmodulated RF signals are applied to the receiver. The baseband noise is measured with a frequency selective voltmeter (FSV) with a measurement bandwidth of 3.1 kilohertz (KHz), the nominal frequency width of a single telephone circuit. The FSV is tuned to different frequencies in the baseband of the receiver while the received signal level (RSL) applied to the receiver is varied. By plotting these noise measurements as a function of baseband frequency and RSL, it is possible, given an actual operational RSL of the receiver, to predict the thermal noise due to the M/W terminal which will be added to a particular multiplexed telephone circuit. Of course, after the noise at the various baseband locations (slot noise) is demultiplexed, the various slot noises will appear in various individual telephone circuits.

1.6 This report deals with the FM noise quieting curve, the various factors which affect it, and the parameters necessary to predict it. The elementary baseband signal characteristics as a function of RSL are also discussed briefly. Due to the wide range of topics covered, this report is limited primarily to formulas and results. Derivations have been limited to those which directly aid the understanding of a concept. Theoretical aspects will not be covered in any depth. This report will briefly cover the important thermal noise characteristics of an FM radio and provide the necessary tools to put that knowledge to use.

2. Basic Modulation Theory

2.1 The purpose of a wideband microwave radio transmission system is to transfer a wide frequency baseband signal from one location to another. Invariably, a sinusoidal radio frequency wave (RF sine wave) is modulated to transfer a baseband signal from one radio terminal to another. Amplitude modulation is a simple modulation process to visualize. A baseband signal is merely transferred to radio frequency. At radio frequency, it may appear with its normal low to high frequency orientation (higher frequency baseband signal producing a higher frequency RF component than a lower frequency baseband signal). In this case, an upper sideband has been produced. If the frequency orientation of the baseband is reversed at RF, then a lower sideband has been produced. Sometimes both sidebands are produced (double sideband), sometimes only one (single sideband). A sinewave (carrier) is used in the modulation process. Sometimes the carrier is eliminated prior to transmission of the RF signal (suppressed carrier). Although there are several forms of amplitude modulation, the processes are similar. Each baseband frequency component produces one or two discrete frequency components in the modulated RF signal. These components are equal in amplitude (except in vestigial sideband modulation) and are separated from the carrier frequency by a frequency difference equal to the frequency of the baseband component. The nature of the modulation is such that the frequency of the RF modulated signal depends only on the carrier frequency and the frequency of the corresponding baseband component. The amplitude of the RF spectral components depend only on the amplitude of the corresponding baseband spectral component's amplitude. The frequency of the modulated signal's spectral components are independent of baseband signal amplitude and vice versa.

2.2 Instead of modulating the amplitude of the transmitted signal, it is possible to transmit baseband signal by changing the angle of the transmitted sinewave relative to its unmodulated condition. The properties of this angle modulation are considerably different than amplitude modulation. Two common methods of achieving angle modulation are called phase modulation (PM) and frequency modulation (FM). The United States Electronic Industry Association (EIA) Standard RS-252-A defines frequency modulation (para 2.1) as "... that process of angle modulation in which the instantaneous frequency deviation of the sinusoidal carrier is proportional to the instantaneous voltage of the modulating signal." The standard defines phase modulation (para 2.2) as "...that process of angle modulation in which the instantaneous phase deviation of the sinusoidal carrier is proportional to the instantaneous voltage of the modulating signal." It should be mentioned that the phase and frequency of a sine wave are related to each other. As Van der Pol has pointed out, the definitions of phase and frequency are not unique. Using Stumper's "zero crossing" definition of frequency has analytical and heuristic advantages and leads to the frequency counter FM demodulators such as those described by Labin. The most common definitions, however, imply that phase is the time integral of frequency and that frequency

is the time derivative of phase. When a carrier is phase modulated, the frequency of the carrier is also modulated. When we specify phase modulation, we are defining a specific relationship between phase changes and the modulating waveform. The frequency will also change, but the relationship between frequency change and the modulation waveform is not explicitly stated. Likewise, when frequency modulation is used, the phase of the carrier is also modulated. All that has been directly defined is a particular relationship between the carrier frequency and the modulating waveform. Using integral and differential calculus relationships, the definition of an equivalent PM (or FM) could be derived for each of the above definitions of FM (or PM). The advantage of the preceeding definitions, however, is their simplicity.

2.3 There are several forms of practical angle modulators and demodulators. Most frequency modulators are voltage controlled oscillators (diode reactance, reactance tube, or klystron type) or zero crossing type (saturable reactor). The diode reactance and klystron type are the most popular for wideband microwave use. Phase modulators are the pulse position type. A popular wideband modulator of this type is the serrasoid. There are no practical wideband phase demodulators. Types of frequency demodulators include the ratio, Foster-Seeley, and Travis detectors, line discriminators, cycle counters, and various forms of phase locked loops and frequency feedback demodulation methods. The Travis discriminator is the most popular type in current high performance wideband terrestrial communication systems.

2.4 Unlike amplitude modulation, the spectrum of the modulated signal (as viewed on a spectrum analyzer, for example) is a complex function of both the frequency and amplitude of the baseband signal. The next page shows the transmit spectrum of an angle modulator for a sine wave baseband signal of different amplitudes. The only difference between the different pictures is the level of the modulating sinewave, yet the spectrums are different. When a single sinewave is used to angle modulate a carrier, a definite relationship exists (as defined by Bessel functions of the first kind of integer order) between the modulation index (beta) of the modulator and the amplitudes of the various sideband components. FM is a complex modulation process. It is not easy to determine from the signal into the modulator what the spectrum of the output will be. In the previously mentioned pictures, the modulated RF spectrum changed considerably, although the input signal did not change its waveform. Certain levels of sine wave modulation cause disappearance of certain RF modulated signal spectral components. Examples of this are given on the next two pages. The relationship between modulator input sine wave level and spectral component disappearance (drop out) will be used later. The voltage and power (dB) amplitude of the spectral components for a sinewave modulated carrier have been tabulated on the next two pages. For comparison, the various spectral components for a square wave frequency modulated carrier are shown on the following two pages. When a single sinewave is the baseband signal

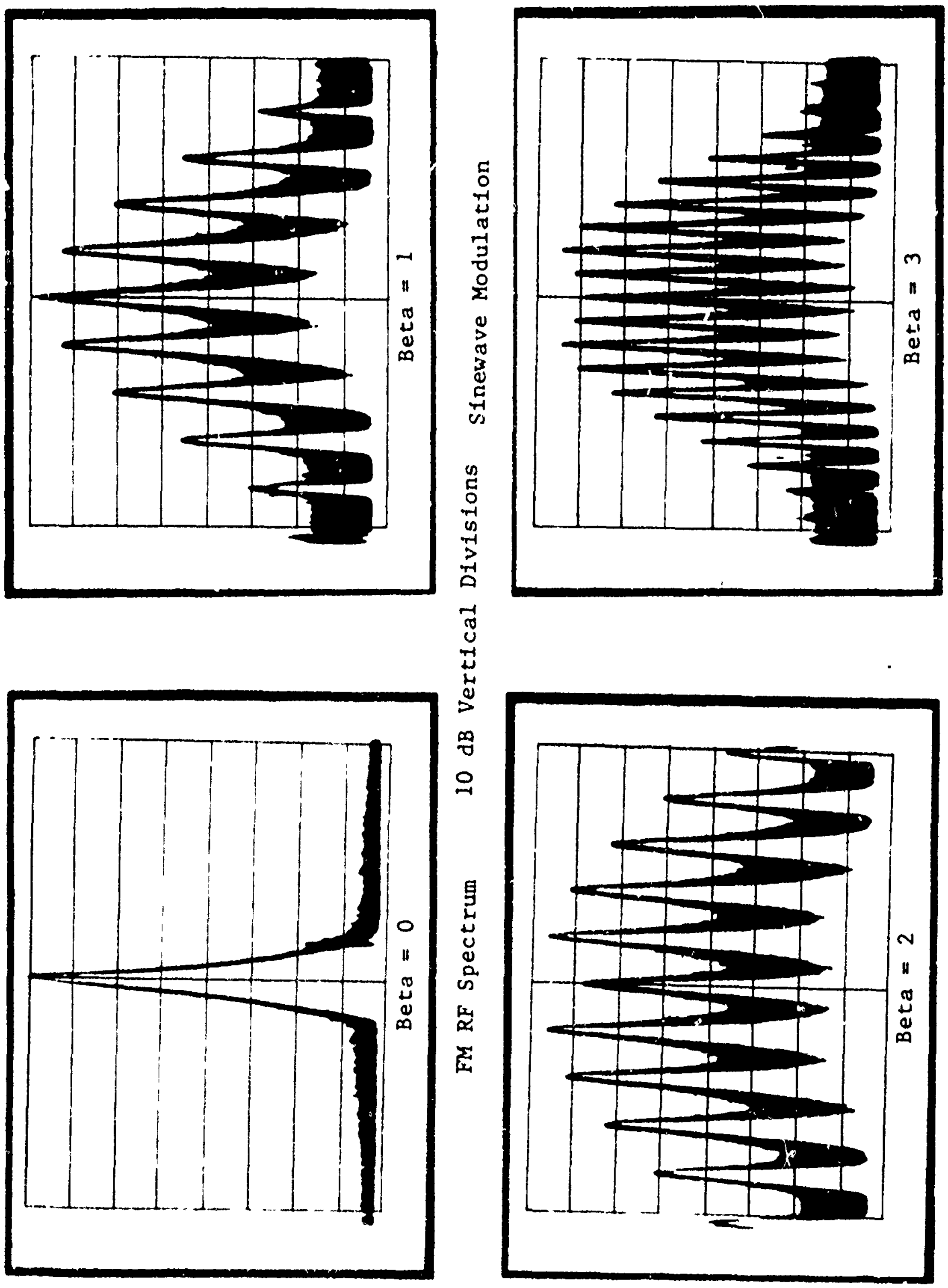
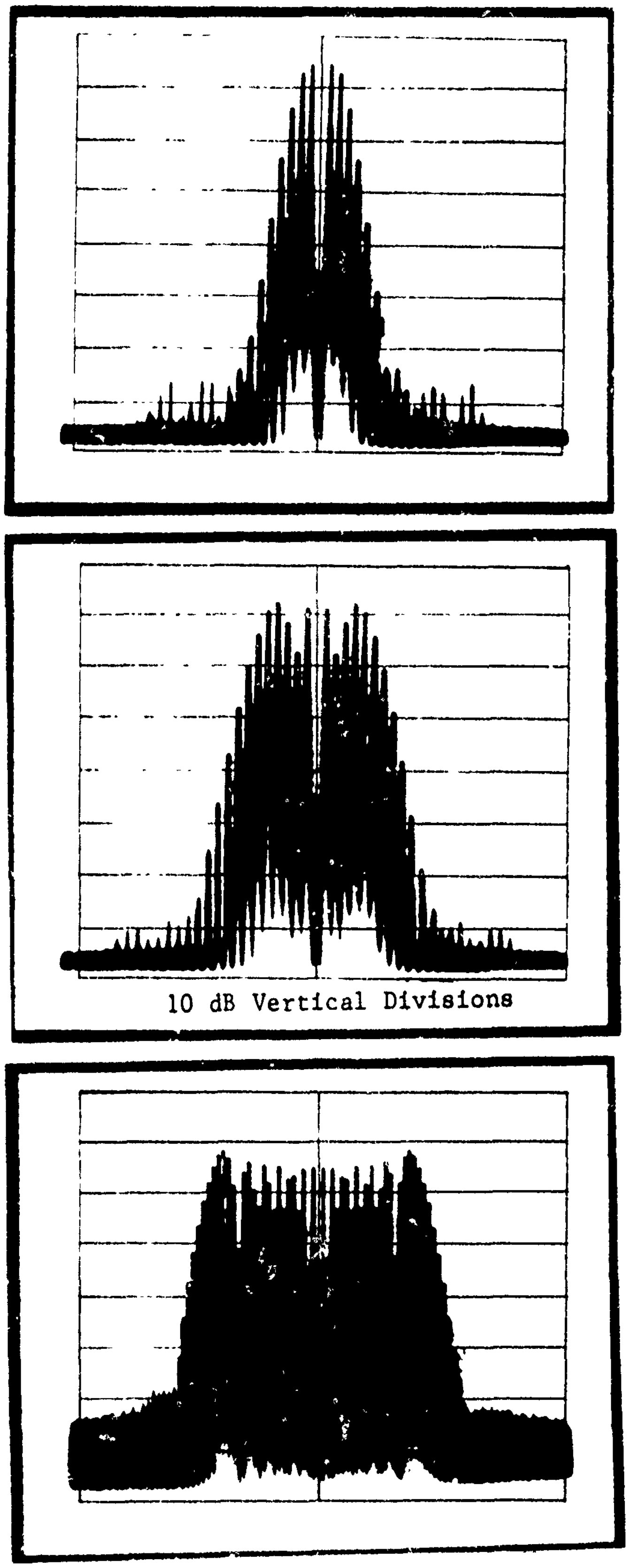

Beta = 1
Beta = 3
Beta = 0
Beta = 2
FM RF Spectrum   10 dB Vertical Divisions   Sinewave Modulation

Figure 1



First Carrier Dropout

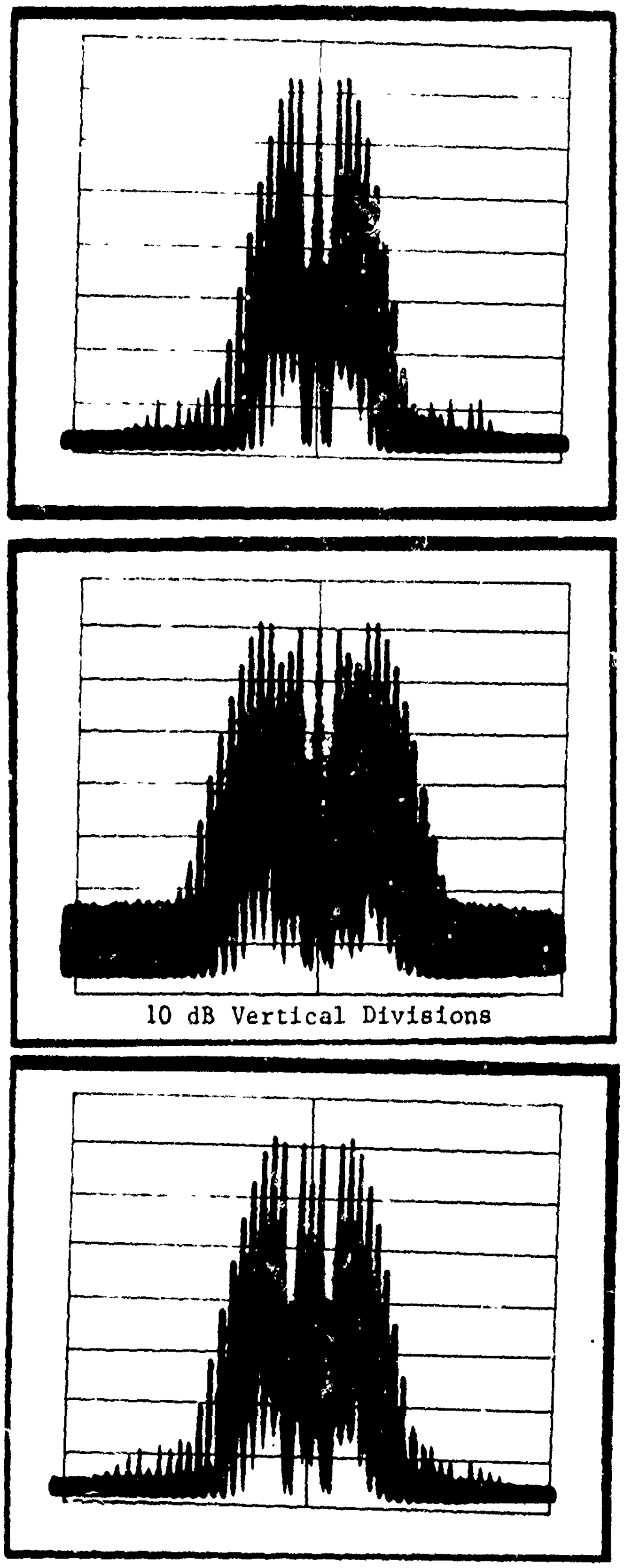
Second Carrier Dropout

Eighth Carrier Dropout

FM RF Spectrum

Sinewave Modulation

Figure 2



First First Sideband Dropout

Second First Sideband Dropout

First Second Sideband Dropout

FM RF Spectrum

Sinewave Modulation

Figure 3

BETA	CARRIER	SIDEBAND 1	SIDEBAND 2	SIDEBAND 3	SIDEBAND 4	SIDEBAND 5
0.	1.0000	0.	0.	0.	0.	0.
0.25	0.9844	0.1240	0.0078	0.0003	0.0000	0.0000
0.50	0.9385	0.2423	0.0306	0.0026	0.0002	0.0000
0.75	0.8642	0.3492	0.0671	0.0085	0.0008	0.0001
1.00	0.7652	0.4401	0.1149	0.0196	0.0025	0.0002
1.25	0.6459	0.5106	0.1711	0.0369	0.0059	0.0007
1.50	0.5118	0.5579	0.2321	0.0610	0.0118	0.0018
1.75	0.3690	0.5802	0.2940	0.0919	0.0209	0.0038
2.00	0.2239	0.5767	0.3528	0.1289	0.0340	0.0070
2.25	0.0827	0.5484	0.4047	0.1711	0.0515	0.0121
2.50	-0.0484	0.4971	0.4461	0.2166	0.0738	0.0195
2.75	-0.1641	0.4260	0.4739	0.2634	0.1007	0.0297
3.00	-0.2601	0.3391	0.4861	0.3091	0.1320	0.0430
3.25	-0.3328	0.2411	0.4811	0.3510	0.1669	0.0599
3.50	-0.3801	0.1374	0.4586	0.3868	0.2044	0.0804
3.75	-0.4014	0.0332	0.4191	0.4138	0.2430	0.1046
4.00	-0.3971	-0.0660	0.3641	0.4302	0.2811	0.1321
4.25	-0.3692	-0.1556	0.2960	0.4341	0.3169	0.1624
4.50	-0.3205	-0.2311	0.2178	0.4247	0.3484	0.1947
4.75	-0.2551	-0.2892	0.1334	0.4015	0.3738	0.2280
5.00	-0.1776	-0.3276	0.0466	0.3648	0.3912	0.2611
5.25	-0.0931	-0.3450	-0.0384	0.3158	0.3993	0.2926
5.50	-0.0068	-0.3414	-0.1173	0.2561	0.3967	0.3209
5.75	0.0760	-0.3179	-0.1866	0.1882	0.3829	0.3446
6.00	0.1506	-0.2767	-0.2429	0.1148	0.3576	0.3621
6.25	0.2131	-0.2207	-0.2837	0.0391	0.3213	0.3721
6.50	0.2601	-0.1538	-0.3074	-0.0353	0.2748	0.3736
6.75	0.2895	-0.0803	-0.3133	-0.1053	0.2196	0.3656
7.00	0.3001	-0.0047	-0.3014	-0.1676	0.1578	0.3479
7.25	0.2920	0.0686	-0.2731	-0.2192	0.0916	0.3204
7.50	0.2663	0.1352	-0.2303	-0.2581	0.0238	0.2835
7.75	0.2252	0.1916	-0.1758	-0.2823	-0.0428	0.2382
8.00	0.1717	0.2346	-0.1130	-0.2911	-0.1054	0.1858
8.25	0.1092	0.2622	-0.0456	-0.2843	-0.1611	0.1281
8.50	0.0419	0.2731	0.0223	-0.2626	-0.2077	0.0671
8.75	-0.0257	0.2672	0.0870	-0.2274	-0.2430	0.0053
9.00	-0.0903	0.2453	0.1448	-0.1809	-0.2655	-0.0550
9.25	-0.1474	0.2091	0.1926	-0.1258	-0.2743	-0.1114
9.50	-0.1939	0.1613	0.2279	-0.0653	-0.2691	-0.1613
9.75	-0.2273	0.1048	0.2488	-0.0028	-0.2505	-0.2028
10.00	-0.2459	0.0435	0.2546	0.0584	-0.2196	-0.2341
10.25	-0.2490	-0.0190	0.2453	0.1147	-0.1781	-0.2537
10.50	-0.2366	-0.0789	0.2216	0.1633	-0.1283	-0.2611
10.75	-0.2101	-0.1325	0.1854	0.2015	-0.0730	-0.2558
11.00	-0.1712	-0.1768	0.1390	0.2274	-0.0150	-0.2383
11.25	-0.1226	-0.2093	0.0854	0.2397	0.0424	-0.2096
11.50	-0.0676	-0.2284	0.0279	0.2381	0.0963	-0.1711
11.75	-0.0096	-0.2331	-0.0300	0.2229	0.1438	-0.1250
12.00	0.0477	-0.2235	-0.0849	0.1951	0.1825	-0.0735
12.25	0.1010	-0.2004	-0.1337	0.1567	0.2104	-0.0193

FREQUENCY MODULATED RF SPECTRUM/SINE WAVE MODULATION INPUT

(levels as voltage ratios)

Table 1

BETA	CARRIER	SIDEBAND 1	SIDEBAND 2	SIDEBAND 3	SIDEBAND 4	SIDEBAND 5
0.	0.	-200.00	-200.00	-200.00	-200.00	-200.00
0.25	-0.14	-18.13	-42.19	-69.78	-99.88	-131.92
0.50	-0.55	-12.31	-30.28	-51.82	-75.88	-101.88
0.75	-1.27	-9.14	-23.47	-41.43	-61.93	-84.38
1.00	-2.32	-7.13	-18.79	-34.17	-52.12	-72.05
1.25	-3.80	-5.84	-15.34	-28.67	-44.62	-62.56
1.50	-5.82	-5.07	-12.69	-24.30	-38.59	-54.90
1.75	-8.66	-4.73	-10.63	-20.74	-33.59	-48.50
2.00	-13.00	-4.78	-9.05	-17.79	-29.37	-43.05
2.25	-21.64	-5.22	-7.86	-15.34	-25.76	-38.33
2.50	-26.31	-6.07	-7.01	-13.29	-22.64	-34.20
2.75	-15.70	-7.41	-6.49	-11.59	-19.94	-30.56
3.00	-11.70	-9.39	-6.27	-10.20	-17.59	-27.32
3.25	-9.56	-12.36	-6.35	-9.09	-15.55	-24.45
3.50	-8.40	-17.24	-6.77	-8.25	-13.79	-21.89
3.75	-7.93	-29.57	-7.55	-7.66	-12.29	-19.61
4.00	-8.02	-23.60	-8.77	-7.33	-11.02	-17.58
4.25	-8.65	-16.16	-10.57	-7.25	-9.98	-15.79
4.50	-9.88	-12.73	-13.24	-7.44	-9.16	-14.21
4.75	-11.87	-10.78	-17.50	-7.93	-8.55	-12.84
5.00	-15.01	-9.69	-26.64	-8.76	-8.15	-11.66
5.25	-20.62	-9.24	-28.32	-10.01	-7.97	-10.67
5.50	-43.29	-9.33	-18.61	-11.83	-8.03	-9.87
5.75	-22.39	-9.95	-14.58	-14.51	-8.34	-9.25
6.00	-16.44	-11.16	-12.29	-18.80	-8.93	-8.82
6.25	-13.43	-13.12	-10.94	-28.15	-9.86	-8.59
6.50	-11.70	-16.26	-10.25	-29.03	-11.22	-8.55
6.75	-10.77	-21.90	-10.08	-19.55	-13.17	-8.74
7.00	-10.46	-46.59	-10.42	-15.52	-16.04	-9.17
7.25	-10.69	-23.28	-11.27	-13.18	-20.76	-9.89
7.50	-11.49	-17.38	-12.76	-11.77	-32.46	-10.95
7.75	-12.95	-14.35	-15.10	-10.98	-27.37	-12.46
8.00	-15.31	-12.59	-18.94	-10.72	-19.55	-14.62
8.25	-19.23	-11.63	-26.81	-10.92	-15.86	-17.85
8.50	-27.55	-11.27	-33.02	-11.61	-13.65	-23.46
8.75	-31.72	-11.46	-21.21	-12.86	-12.29	-45.55
9.00	-20.88	-12.21	-16.78	-14.85	-11.52	-25.19
9.25	-16.63	-13.59	-14.31	-18.00	-11.24	-19.07
9.50	-14.25	-15.85	-12.85	-23.70	-11.40	-15.85
9.75	-12.87	-19.59	-12.08	-51.21	-12.02	-13.86
10.00	-12.18	-27.24	-11.88	-24.68	-13.17	-12.61
10.25	-12.08	-34.42	-12.21	-18.81	-14.99	-11.91
10.50	-12.52	-22.06	-13.09	-15.74	-17.83	-11.67
10.75	-13.55	-17.56	-14.64	-13.92	-22.74	-11.84
11.00	-15.33	-15.05	-17.14	-12.87	-36.46	-12.46
11.25	-18.23	-13.58	-21.37	-12.41	-27.45	-13.57
11.50	-23.40	-12.83	-31.08	-12.47	-20.33	-15.33
11.75	-40.34	-12.65	-30.45	-13.04	-16.84	-18.06
12.00	-26.42	-13.02	-21.42	-14.19	-14.77	-22.68
12.25	-19.92	-13.96	-17.48	-16.10	-13.54	-34.29

FREQUENCY MODULATED RF SPECTRUM/SINE WAVE MODULATION INPUT

(levels in dB)

Table 2

BETA	CARRIER	SIDEBAND 1	SIDEBAND 2	SIDEBAND 3	SIDEBAND 4	SIDEBAND 5
0.	1.0000	0.	0.	0.	0.	0.
0.25	0.9745	0.1568	0.0155	-0.0165	-0.0038	0.0059
0.50	0.9003	0.3001	0.0600	-0.0257	-0.0143	0.0091
0.75	0.7842	0.4176	0.1283	-0.0217	-0.0286	0.0075
1.00	0.6366	0.5000	0.2122	0.0000	-0.0424	-0.0000
1.25	0.4705	0.5414	0.3016	0.0409	-0.0509	-0.0130
1.50	0.3001	0.5402	0.3858	0.1000	-0.0491	-0.0297
1.75	0.1392	0.4990	0.4548	0.1734	-0.0330	-0.0469
2.00	0.0000	0.4244	0.5000	0.2546	0.0000	-0.0606
2.25	-0.1083	0.3258	0.5159	0.3361	0.0501	-0.0664
2.50	-0.1801	0.2144	0.5002	0.4092	0.1154	-0.0600
2.75	-0.2139	0.1021	0.4540	0.4661	0.1917	-0.0384
3.00	-0.2122	0.0000	0.3820	0.5000	0.2728	0.0000
3.25	-0.1810	-0.0828	0.2913	0.5067	0.3515	0.0548
3.50	-0.1286	-0.1400	0.1910	0.4848	0.4201	0.1236
3.75	-0.0650	-0.1688	0.0908	0.4357	0.4715	0.2017
4.00	-0.0000	-0.1698	0.0000	0.3638	0.5000	0.2829
4.25	0.0573	-0.1465	-0.0736	0.2756	0.5020	0.3603
4.50	0.1000	-0.1052	-0.1247	0.1801	0.4766	0.4265
4.75	0.1238	-0.0537	-0.1505	0.0853	0.4257	0.4748
5.00	0.1273	-0.0000	-0.1516	0.0000	0.3537	0.5000
5.25	0.1120	0.0482	-0.1310	-0.0689	0.2671	0.4991
5.50	0.0818	0.0846	-0.0943	-0.1165	0.1737	0.4716
5.75	0.0424	0.1055	-0.0482	-0.1405	0.0821	0.4195
6.00	0.0000	0.1091	-0.0000	-0.1415	0.0000	0.3472
6.25	-0.0390	0.0966	0.0434	-0.1223	-0.0660	0.2614
6.50	-0.0693	0.0709	0.0765	-0.0880	-0.1115	0.1696
6.75	-0.0871	0.0369	0.0955	-0.0450	-0.1343	0.0800
7.00	-0.0909	0.0000	0.0990	-0.0000	-0.1350	0.0000
7.25	-0.0811	-0.0343	0.0878	0.0405	-0.1166	-0.0641
7.50	-0.0600	-0.0611	0.0646	0.0715	-0.0839	-0.1080
7.75	-0.0314	-0.0772	0.0337	0.0893	-0.0429	-0.1300
8.00	-0.0000	-0.0808	0.0000	0.0926	-0.0000	-0.1306
8.25	0.0295	-0.0724	-0.0314	0.0822	0.0386	-0.1127
8.50	0.0530	-0.0537	-0.0561	0.0605	0.0680	-0.0810
8.75	0.0672	-0.0282	-0.0709	0.0316	0.0850	-0.0413
9.00	0.0707	-0.0000	-0.0744	0.0000	0.0881	-0.0000
9.25	0.0636	0.0266	-0.0667	-0.0294	0.0782	0.0372
9.50	0.0474	0.0479	-0.0496	-0.0526	0.0576	0.0655
9.75	0.0250	0.0610	-0.0261	-0.0666	0.0300	0.0818
10.00	0.0000	0.0643	-0.0000	-0.0700	0.0000	0.0849
10.25	-0.0238	0.0579	0.0247	-0.0628	-0.0280	0.0753
10.50	-0.0429	0.0433	0.0445	-0.0467	-0.0502	0.0554
10.75	-0.0547	0.0229	0.0567	-0.0246	-0.0635	0.0289
11.00	-0.0579	0.0000	0.0599	-0.0000	-0.0667	0.0000
11.25	-0.0523	-0.0218	0.0540	0.0233	-0.0598	-0.0270
11.50	-0.0391	-0.0394	0.0404	0.0420	-0.0445	-0.0483
11.75	-0.0207	-0.0504	0.0214	0.0535	-0.0235	-0.0611
12.00	-0.0000	-0.0534	0.0000	0.0566	-0.0000	-0.0642
12.25	0.0199	-0.0483	-0.0204	0.0511	0.0223	-0.0576

FREQUENCY MODULATED RF SPECTRUM/SQUARE WAVE MODULATION INPUT

(levels as voltage ratios)

Table 3

BETA	CARRIER	SIDEBAND 1	SIDEBAND 2	SIDEBAND 3	SIDEBAND 4	SIDEBAND 5
0.	0.	-200.00	-200.00	-200.00	-200.00	-200.00
0.25	-0.22	-16.09	-36.21	-35.68	-48.36	-44.59
0.50	-0.91	-10.45	-24.43	-31.79	-36.90	-40.82
0.75	-2.11	-7.58	-17.83	-33.29	-30.88	-42.53
1.00	-3.92	-6.02	-13.46	-166.19	-27.44	-169.72
1.25	-6.55	-5.33	-10.41	-27.76	-25.86	-37.73
1.50	-10.45	-5.35	-8.27	-20.00	-26.18	-30.55
1.75	-17.13	-6.04	-6.84	-15.22	-29.64	-26.57
2.00	-154.15	-7.44	-6.02	-11.88	-163.70	-24.35
2.25	-19.31	-9.74	-5.75	-9.47	-26.00	-23.56
2.50	-14.89	-13.38	-6.02	-7.76	-18.75	-24.43
2.75	-13.40	-19.82	-6.86	-6.63	-14.35	-28.31
3.00	-13.46	-156.65	-8.36	-6.02	-11.28	-162.67
3.25	-14.85	-21.64	-10.71	-5.90	-9.08	-25.22
3.50	-17.81	-17.07	-14.38	-6.29	-7.53	-18.16
3.75	-23.75	-15.45	-20.84	-7.22	-6.53	-13.91
4.00	-154.15	-15.40	-157.67	-8.78	-6.02	-10.97
4.25	-24.83	-16.68	-22.66	-11.19	-5.99	-8.87
4.50	-20.00	-19.56	-18.09	-14.89	-6.44	-7.40
4.75	-18.14	-25.41	-16.45	-21.38	-7.42	-6.47
5.00	-17.90	-155.74	-16.39	-158.24	-9.03	-6.02
5.25	-19.01	-26.35	-17.65	-23.24	-11.47	-6.04
5.50	-21.74	-21.45	-20.51	-18.67	-15.20	-6.53
5.75	-27.46	-19.54	-26.34	-17.04	-21.71	-7.55
6.00	-152.85	-19.24	-156.65	-16.99	-158.59	-9.19
6.25	-28.18	-20.30	-27.24	-18.25	-23.61	-11.65
6.50	-23.19	-22.98	-22.33	-21.11	-19.06	-15.41
6.75	-21.20	-28.66	-20.40	-26.94	-17.44	-21.94
7.00	-20.82	-154.01	-20.08	-157.25	-17.39	-158.83
7.25	-21.82	-29.31	-21.13	-27.84	-18.66	-23.87
7.50	-24.43	-24.28	-23.79	-22.92	-21.53	-19.33
7.75	-30.05	-22.25	-29.45	-20.99	-27.36	-17.72
8.00	-154.15	-21.85	-154.79	-20.67	-157.67	-17.68
8.25	-30.59	-22.81	-30.07	-21.71	-28.27	-18.96
8.50	-25.52	-25.40	-25.03	-24.37	-23.35	-21.83
8.75	-23.45	-30.99	-22.98	-30.02	-21.41	-27.67
9.00	-23.01	-155.07	-22.57	-155.35	-21.10	-157.99
9.25	-23.93	-31.49	-23.52	-30.62	-22.13	-28.59
9.50	-26.49	-26.39	-26.09	-25.57	-24.79	-23.67
9.75	-32.05	-24.30	-31.67	-23.53	-30.44	-21.74
10.00	-155.04	-23.84	-155.74	-23.10	-155.78	-21.42
10.25	-32.48	-24.74	-32.14	-24.05	-31.05	-22.46
10.50	-27.36	-27.28	-27.04	-26.62	-25.99	-25.12
10.75	-25.24	-32.82	-24.93	-32.19	-23.94	-30.78
11.00	-24.75	-155.79	-24.46	-156.25	-23.52	-156.11
11.25	-25.63	-33.22	-25.35	-32.65	-24.46	-31.38
11.50	-28.15	-28.08	-27.88	-27.53	-27.03	-26.33
11.75	-33.67	-25.95	-33.41	-25.43	-32.60	-24.28
12.00	-152.85	-25.45	-156.38	-24.95	-156.65	-23.85
12.25	-34.03	-26.31	-33.79	-25.84	-33.05	-24.79

FREQUENCY MODULATED RF SPECTRUM/SQUARE WAVE MODULATION INPUT
(levels in dB)

Table 4

to a modulator, the frequency separation between the various spectral components of the modulated signal is always equal to the sinewave frequency. When two or more sinewave baseband signals are used simultaneously, the frequencies of the sideband components are separated from the carrier by all possible combinations of frequencies which can be obtained from sum and differences of all harmonics of the modulating sinewaves. For many possible wideband baseband signals, an exact determination of the transmit signal spectrum is not practical. About the only practical method of analysis of the baseband/transmit signal relationship is on a statistical basis. There is, however, a general relationship between the modulating signal and the modulated waveform. This generalization is mentioned by both Middleton and Giacoletto. Middleton calls it the "Principle of Adiabatic Frequency Sweeps". The principle simply states that for any general modulation waveform producing frequency modulation, the transmitted radio frequency (RF) spectrum will be symmetric about the unmodulated carrier frequency if the modulating signal has voltages which, when viewed in time, are symmetric about zero volts. Conversely, unsymmetrical modulating signals produce unsymmetrical FM modulated RF spectrums. This is a useful principle. It indicates that if an FM modulator is driven by a sinewave, the RF spectrum will be symmetric. If it is not, the modulated wave has unquestionably been distorted. The FM receiver requires a symmetric received RF spectrum to produce a symmetric baseband signal like a sinewave. It receives an unsymmetric RF spectrum, it can only produce an unsymmetric voltage waveform at the baseband demodulator output. If a sinewave (a symmetric waveform) is transmitted and an unsymmetric waveform is received, the baseband signal has been distorted.

2.5 There are several ways that RF spectrum can be made unsymmetric. At high baseband modulator input levels, nonlinearity of the baseband amplifiers and the modulator itself are distinct possibilities. However, even at low drive levels, it is possible to make the RF spectrum unsymmetric. As Middleton and Panter have mentioned, simultaneous (correlated) amplitude modulation and frequency modulation invariably produce an unsymmetric RF spectrum. It is possible to have this take place right at the FM modulator. Most FM modulators are immediately followed by hard limiters to avoid this possibility. If the RF circuitry has nonuniform amplitude frequency response, or a uniform but unsymmetric (about the carrier frequency) frequency response, correlated AM and PM will be introduced into the signal. In severe cases, either of these can cause distortion of the normal sideband amplitudes and, therefore, distortion in the demodulated signal.

2.6 The modulation index of an angle modulated signal is a factor which relates the modulating waveform to the phase deviation of the modulated signal. This factor is called beta (B) and has the units radians per millivolt. For most baseband signals, a beta is difficult or impossible to define. However, for a sinewave baseband input, the determination of beta is straightforward. For a phase modulation system, the beta is just the specified phase sensitivity of the modulator. For a frequency

modulation system, beta is the specified frequency deviation sensitivity of the modulator (kilohertz per millivolt) divided by the frequency (kilohertz) of the modulating sinewave. The division of frequency deviation by frequency is a direct result of having to integrate the sinusoidal frequency deviation to obtain equivalent phase deviation. As mentioned earlier, knowing the sinewave beta of the angle modulation transmitter allows us to predict exactly (using Bessel functions) the RF transmit spectrum for sinewave modulation.

2.7 Like amplitude modulation, there are many possible variations of angle modulation. As mentioned previously, however, there are two basic angle modulation types. With amplitude modulation, it is possible, with suitable processing of the signals, to turn one type of modulation into another (for example, suppress the carrier, filter one sideband, etc.). As with amplitude modulation, (see, for example, Black and Taub and Schilling.) with suitable processing a phase modulation system can be converted into a frequency modulation system and vice versa. Specifically, if the baseband input of a phase modulator is preceeded by a time integrating circuit, the combination will have all the properties of a frequency modulator. Likewise, if the baseband output of a phase demodulator is coupled to a time differentiation circuit, the combination will have all the properties of a frequency demodulator. A similar procedure holds for frequency modulation. A frequency modulator with baseband input preceeded by a time differentiation circuit has exactly the same properties as a phase modulator. A frequency demodulator whose baseband output is followed by a time integration network has exactly the same properties as a phase demodulator. Therefore, it is theoretically possible to mix and match various modulators and demodulators to achieve a desired microwave radio system. These properties are not theoretical abstractions. The Radio Equipment Laboratories (REL) military tropospheric scatter (TROPO) microwave radio system AN/FRC-39(V) is basically a phase modulation, frequency demodulation system. The baseband circuitry of the modulator and demodulator is such that overall the modulator and demodulator are completely compatible. In fact, the serrisoid phase modulator used in the AN/FRC-39(V), when preceeded by the baseband corrector network, produces a transmit RF signal which is completely indistinguishable from a comparable FM modulator preceeding by the appropriate pre-emphasis network.

2.8 Frequency and phase demodulators have different noise properties. A strong unmodulated carrier centered in the symetric (IF) passband of a phase demodulator produces a flat (white) spectrum of thermal noise over the entire output baseband if the only noise generated in the receiver is white thermal noise. In frequency division multiplexing (FDM) used to produce an analog baseband to apply to an angle modulator, different 3.1 kilohertz wide telephone channels appear at different 3.1 kilohertz frequency positions (slots) in the radio baseband. Therefore, for strong received signal levels, a phase demodulator will introduce the same noise in all telephone slots. When the telephone channels are demultiplexed, the radio noise will be the same for all channels.

With a frequency demodulator under the same conditions, the thermal noise voltage in the receive baseband increases linearly with frequency (hence, it is often called triangular noise). Since power is proportional to the voltage squared, the noise power increases with the square of the baseband (slot) frequency. Therefore, the noise power increases at a rate of 6 dB per octave or 20 dB per decade. When there is no carrier present at the receiver, the noise spectrum (as Crosby and Downing note, for example) is essentially the same (flat) for all baseband frequencies (so called rectangular noise). Under the same conditions, (as noted by Taub and Schilling, for example) the noise voltage in the baseband of the phase demodulator decreases inversely with baseband slot frequency. Therefore, the noise power is reduced 6 dB per octave (20 dB per decade). Due to the flat noise spectrum in the baseband of a phase demodulator under strong received signal conditions, it is a highly desirable type of angle modulation receiver. High performance wideband phase demodulators are difficult to construct. At the transmitter, the problems are more difficult. Most phase modulators can only produce low modulation index modulation. To obtain the necessary modulation index with a phase modulator, multiplication of the modulator output is generally required. Using a multiplier puts some severe requirements on the basic phase noise performance of the modulator's oscillator. Every time the oscillator output is doubled, the noise it produces at the baseband of a receiver with a strong RSL (receiver in saturated region) is increased 6 dB. From a practical point of view, it is much easier to use a voltage controlled oscillator as the modulation source. With such modulators, wide linear frequency shifts directly proportional to input voltage can be obtained. Little or no multiplication is required, and the modulated signal can often be produced directly at RF without the need of an up converter. Using frequency modulators, high performance wideband microwave systems are readily achievable.

2.9 If we wish to use a phase demodulator, we must turn the frequency modulator into a phase modulator to be compatible with a phase demodulator. To turn a frequency modulator into a phase modulator, the baseband input to the modulator must be preceded by a differentiation network. Taking advantage of the elementary properties of Fourier transforms, if a device causes differentiation in the time domain, it will cause an additional term directly proportional to frequency in the voltage versus frequency response of the overall system. Since power is proportional to voltage squared, the power frequency response of a baseband circuit with a differentiator in it will have a frequency response which increases with the square of frequency. Again, this is 6 dB per octave. This is not an undesirable feature for the telephone channels in the high end of the baseband; their level is boosted through the differentiation network. However, relative to the high end of the baseband, the telephone channels at the low end of the baseband will be suppressed. For a 600 telephone channel radio baseband extending from 60 kilohertz to 2660 kilohertz, for example, out of the differentiation network the highest telephone

channel would be 33 dB greater in level than the lowest telephone channel. To recover the baseband at the receive end, the integrator network at the frequency demodulator must boost the lowest channel 33 dB relative to the highest telephone channel so that uniform frequency response is maintained in and out of the radio terminal. In a noiseless transmission system, attenuating the low end of the baseband 33 dB and then boosting it 33 dB at the receive baseband would be no problem. Unfortunately, all transmitters use oscillators which have significant phase jitter. This jitter causes a roughly flat noise spectrum to be produced at the baseband of a companion frequency demodulator (if the received signal level is sufficiently high that thermal noise is negligible compared to the transmit phase noise). Because of the significant amount of transmit phase noise (Strictly speaking, the noise is equivalent frequency noise produced by phase noise.) that will appear at the baseband of the FM receiver, attenuating this level of the low frequency telephone channels by many dB is out of the question. The signal to noise ratio of the low frequency channels will have been excessively degraded at the receiver by the transmitters phase noise. Because of these considerations, actual differentiation and integration networks are never used with frequency modulation modulators and demodulators. Instead, devices called pre-emphasis and de-emphasis networks are used. These networks have essentially the frequency response of differentiators and integrators over about the top octave of the radio baseband. Over the rest of the baseband, the networks have essentially constant frequency response. Frequency modulation radio terminals with pre-emphasis and de-emphasis are actually a hybrid of frequency and phase modulation systems. Over about the top octave of the radio baseband, they have noise characteristics of phase modulation systems. Over the rest of the baseband, however, they have the noise characteristics of a frequency modulation system. During the rest of this report, frequency modulation (FM) systems will be considered exclusively. However, it should be kept in mind that when FM systems use pre-emphasis and de-emphasis networks, they achieve some of the thermal noise properties of phase modulation systems. The effect of de-emphasis networks on thermal noise performance of FM receivers will be included in that report.

2.10 At this point, the character of baseband noise has only been inferred for high carrier to thermal noise ratios and for the case where there is no received signal. As the carrier level of the received microwave signal is reduced from a strong RSL, the noise power in the baseband of the receiver increases. As the carrier level at the receiver is reduced 1 dB, the noise at the baseband output of the receiver increases 1 dB. The frequency distribution (shape) of the noise remains the same. If viewed on an oscilliscope, the noise would appear as the usual relatively smooth looking background thermal noise. As the received signal level approaches low levels, the character of the noise starts to change. In addition to the smooth noise, noise spikes (impulses) begin to appear. As the received signal gets lower and lower, the individual noise spikes or clicks occur more and more often. With continued reduction in received signal, the clicks rapidly merge into a crackling or sputtering noise.

Finally, the clicks murge to produce a continuous high level of noise out of the baseband of the receiver. As the noise impulses appear more and more often, the noise in the various baseband slots starts to increase faster than 1 dB for 1 dB decrease in received signal level. The received signal level at which the noise in a narrow measurement slot in the baseband increases 2 dB for a 1 dB reduction in carrier level is generally called FM noise threshold. Since the impulse noise is essentially flat in power spectrum, the noise increase is first seen in the low baseband frequencies where the normal thermal noise is quite low. The difference in FM threshold between the highest and lowest slot in a wideband receiver can be several dB. For standardization, FM noise threshold is normally measured in one of the highest baseband slots. For those interested in why impulses occur for low received signal levels, Taub and Schilling's text is highly recommended. In addition to the basic thermal noise properties of FM systems, there are several other properties of interest.

2.11 Because FM signals carry no information in the amplitude of the transmitted signal, it is possible to use a limiter in the receiver to suppress any residual amplitude modulation in the received signal before it is demodulated. The use of a limiter in an FM receiver can introduce additional intermodulation when receive signals have passed through certain types of amplitude nonlinearities. Middleton has pointed out that not using a limiter, however, leaves the signal prone to another type of intermodulation distortion. In general limiters are used for a couple of very good reasons. The received signal and thermal noise performance of the receiver is significantly improved near FM noise threshold when the limiter is used. For large received signal to thermal noise (large carrier to noise ratio) conditions (i.e., well above FM noise threshold), the use of heavy limiting gives the FM receiver its ability to reject unwanted signals lower in power than the desired signal. This ability is often expressed as capture ratio. This ability is very important to high quality microwave communication since it is directly responsible for the ability of the receiver to reject other signals on its frequency (co-channel interference), signals on frequencies slightly different than its own (adjacent channel interference), and its own signal reflected from stray sources (multipath). Since all wideband microwave receivers use hard limiters (heavy limiting), many authors have stated or implied that FM receivers have an inherent ability to reject other signals of lower received power level. This simply is not true. Without hard limiting action, the FM receiver loses much, if not all, of its ability to reject unwanted signals. Downing specifically mentions that the ability of an FM receiver to reject impulse noise and coherent interference is directly attributable to the receivers amplitude limiter. Corrington mentions an interesting example. During some propagation tests he was conducting, he observed two distant FM stations. Both stations were on the same frequency and of nearly equal receive signal strength. However, each signal was randomly fading up and down in received signal level. His receiver was such that it achieved limiting for strong received signals, but did not limit for weak received signals. When both signals were strong, the limiter was saturated and

only the stronger signal's program was heard. When the other signal became stronger, there would be a short burst of noise and the program would abruptly change to that of the stronger signal. For many minutes, conditions were such that the programs changed back and forth about once every fifteen seconds. Occasionally, both signal levels fell below the level at which the receiver limited. During those periods of time, both programs could be heard simultaneously.

2.12 There are two important considerations to insure hard limiting. Panter mentions that in order to achieve high suppression of unwanted interference, the limiter and the stages preceding the first limiter stage must have essentially flat frequency response over the frequency range of the desired signal. Ruthroff mentions the other important consideration. It is well understood that in limiter circuits, a minimum level into the limiter is necessary to achieve limiting. Contrary to common belief, however, merely increasing the drive level does not necessarily improve limiting. Strangely enough, overdriving a practical limiter circuit can seriously degrade it. For circuits of finite frequency bandwidth and utilizing limiter diodes with finite back resistance, there is some unique input level at which limiting is best. The purpose of Automatic Gain Control (AGC) in an FM receiver is to keep the carrier input level near the value for best limiting.

2.13 One other factor in good interference rejection is the bandwidth and symmetry of the FM demodulator (discriminator). As Panter mentions, for best capture ratio the discriminator must be able to accommodate the necessary amplitude and frequency excursions which occur in an interference situation. Leentvaar and Flint have shown that it is possible to reduce or eliminate the capture effect of an FM receiver by reducing the bandwidth of an FM demodulator.

2.14 Due to the nonlinear demodulation action of an FM receiver, limiters and the general nonlinear process of FM generation and detection, the noise produced in an FM receiver due to an arbitrary interfering signal is difficult to predict with any degree of precision. The articles listed in the bibliography are available for information on specific interference situations. A few generalizations, however, are available. Corrington suggests some effects due to a single unmodulated interfering signal. Introducing an unwanted carrier into an FM receiver increases the overall noise demodulated at the baseband of the receiver. At a low level relative to the desired signal (high carrier to interference ratio) the increased noise in the baseband of the FM receiver is primarily at low baseband frequencies. As the interfering signal gets stronger, the noise at higher baseband frequencies increases faster than the noise at lower frequencies. When the desired signal and the interference are about the same level (C/I= 0dB), the baseband noise is quite high and essentially the same at all baseband frequencies.

2.15 If both the desired and the interfering signal are unmodulated or only lightly modulated (loaded) FDMed FM signals, a high density, very narrow noise spike is produced in the receiver baseband. This spike is a crossmodulation (beat) product caused by the modulation of the desired signal by the interference. The baseband frequency of the noise spike is the difference between the frequency of the unmodulated signal and the desired signal. If the interfering signal or the main signal is highly loaded but the other is lightly loaded, noise is produced in the baseband and is concentrated around the baseband frequency corresponding to the difference between the center (carrier) frequency of the desired and the interfering signals. If both signals are highly modulated, broadband interference can be expected. Unlike the case when the signals are unmodulated or only lightly modulated, the noise appears as broadband intermodulation noise.

2.16 In addition to interference considerations, Taub and Schilling and Panter mention that if the received signal is not centered in the center of the IF of the receiver or if the IF frequency response is not symmetric with respect to the normal center frequency, the noise out of the baseband will be greater than normal. Rice, Sundie, and Taub and Schilling indicate that the FM threshold of an FM receiver will be degraded due to the production of additional baseband noise if the received signal is modulated. Middleton indicates if an FM modulator is wideband white noise modulated and the signal is demodulated by a receiver with no limiter, the receive baseband will experience considerable essentially flat noise even for strong received signal levels.

2.17 One other noise characteristic of FM receivers is the noise out of the receiver at received signal levels at FM threshold and less. As Middleton has shown, the character of the noise out of the baseband of an FM receiver changes dramatically as the limiting in an FM receiver is changed from hard to soft to no limiting. If limiting action is reduced or lost in the region of FM threshold, the noise in the baseband increases dramatically. The baseband noise is basically wide frequency range flat spectrum noise which can significantly degrade FM threshold. It is interesting, however, that with no carrier present at the receiver and no receiver limiting, the noise out of the baseband receiver is invariably less than the noise that would occur with heavy limiting. The noise out of the baseband of the radio with no carrier present is a direct function of the amount of limiting in the receiver. The relation between amount of limiting and baseband noise is highly nonlinear, however. Several dB of baseband noise is lost as limiting transitions from hard to soft.

2.18 To this point, no mention has been made as to how the baseband signal is affected as the received carrier is reduced into the region near FM threshold. With hard receiver limiting, the baseband signal remains constant. However, as the carrier falls below the FM threshold level, the baseband signal is suppressed rather abruptly. Below FM

threshold, the demodulated baseband signal is reduced 2 dB for every dB of reduced signal level. As with baseband noise near FM threshold, the limiter affects the suppression of the baseband signal. The best performance is obtained with hard limiting. As limiting goes soft, the signal suppression starts earlier. Middleton has shown that the effect with soft limiting is to cause the baseband signal level to start being lost at a much stronger received signal level. The baseband signal level is lost more gradually. With soft limiting, the received baseband signal level eventually drops off at the same 2 dB per 1 dB of received signal reduction as occurred for hard limiting. However, with soft limiting, a given amount of signal suppression will occur at a stronger received signal level.

3. Wideband FM Receiver Baseband Signal and Noise Characteristics

3.1 If the noise in a single baseband noise slot were plotted as a function of RF Received Signal Level (RSL), the plot would appear as shown on the next page.

3.2 Region A is the region of the quieting curve where the RF signal has essentially been lost and the slot noise is dependent on the thermal noise generated within the front end of the receiver, the receiver IF response, and the amount of receiver limiting. Region B is the non-linear region of the quieting curve. This region is a complex function of the signal introduced into the receiver, the thermal noise generated, the receiver IF response, and degree of signal limiting. Region C is the linear region of the receiver. Here the noise decreases in direct proportion to the increase in signal introduced into the receiver. Region D is the so-called saturated region of the receiver. Here, the slot noise at the baseband of the radio is independent of the received signal level and receiver generated noise.

3.3 Describing the noise in regions A, and B is a complex problem. Fortunately, the problem has been solved by several people (e.g. Middleton, Rice, Stumpers, and Wang). Most of the solutions are quite complicated. The solutions involve multiple infinite series and cross products of confluent hypergeometric functions or multiple indefinite integrals of complicated infinite series. Taking advantage of the previous solutions to the problem, it is possible to tabulate the slot noise for various carrier to thermal noise (C/N) ratios (assuming hard limiting). The tabulated results on the next few pages have been normalized. To convert the tabulated noise values to dBmØ noise in a 3.1 kHz baseband slot, add the following factor:

$$No\ (dBm\emptyset) = +29.8 - 20 \log \Delta f_{/ch\ rms} + 10 \log B - P$$

where

$\Delta f_{/ch\ rms}$ = per channel rms deviation in KHz

B = receiver IF bandwidth in MHz

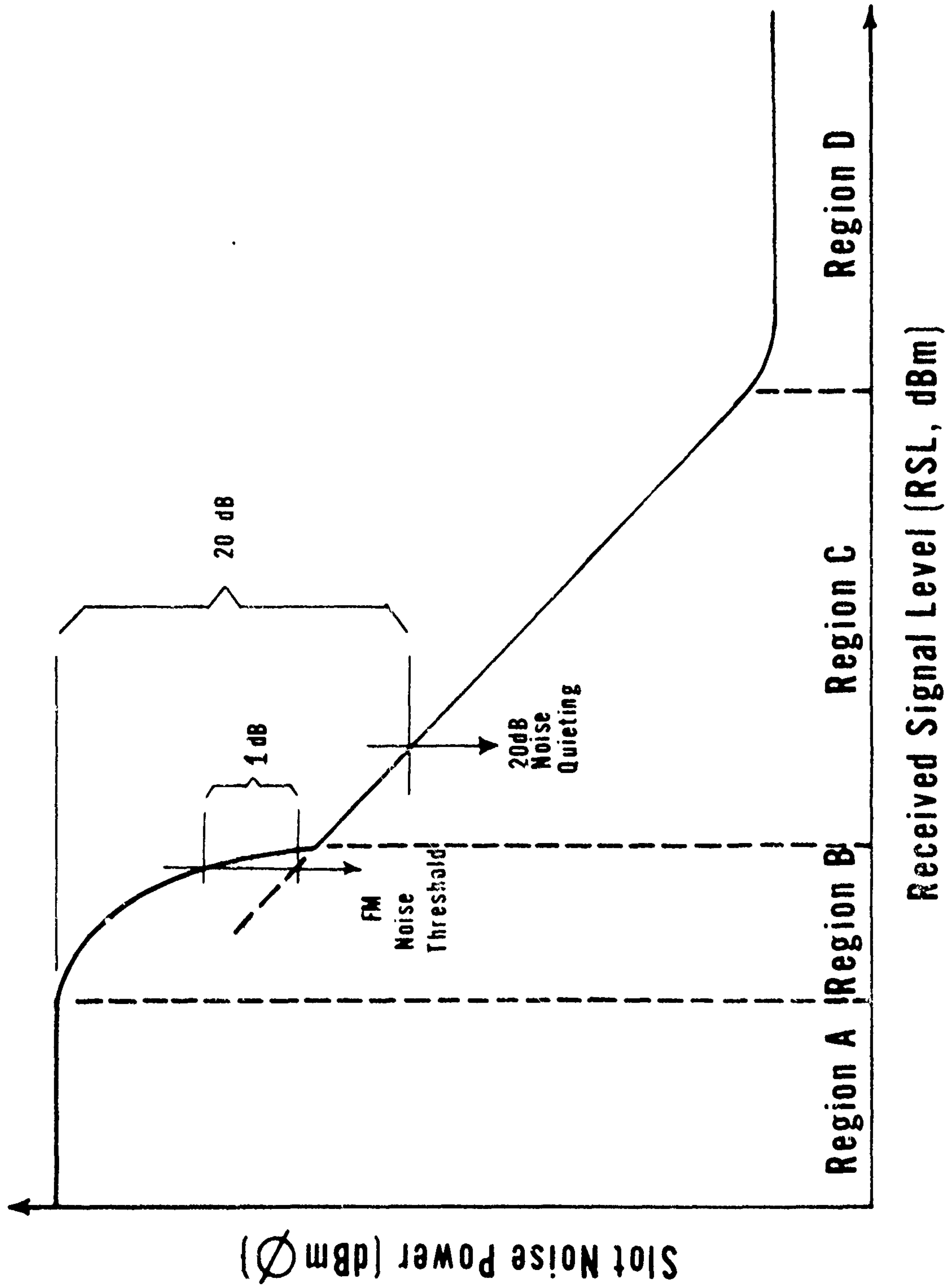
P = baseband pre-emphasis in dB relative to baseband pivot frequency (formulas for this are given later in this report)

To convert the C/N values to RSL, use the following formula:

$$RSL(dBm) = C/N(dB) - 114.0 + NF + 10 \log B$$

where

NF = receiver overall noise figure in dB measured at the same point at which RSL is to be measured



Slot Noise Versus Received Signal Level (RSL)

Figure 4

	C/N (dB)	C/N (PR)	Diff. (dB)	Rect. Noise (dB)	Gaus. Noise (dB)
F/B=0.2					
	5.	3.162	0.6	-11.7	-11.1
	4.	2.512	1.2	-10.8	-9.5
	3.	1.995	1.6	-9.6	-8.0
	2.	1.585	1.8	-8.4	-6.7
	1.	1.259	1.9	-7.3	-5.5
	0.	1.000	1.9	-6.4	-4.5
	-1.	0.794	1.9	-5.5	-3.6
	-2.	0.631	1.9	-4.8	-2.9
	-3.	0.501	1.9	-4.2	-2.4
	-4.	0.398	1.8	-3.7	-1.9
	-5.	0.316	1.8	-3.3	-1.5
	-6.	0.251	1.8	-3.0	-1.2
	-7.	0.200	1.8	-2.8	-1.0
	-8.	0.158	1.7	-2.5	-0.8
	-9.	0.126	1.7	-2.4	-0.6
	-10.	0.100	1.7	-2.2	-0.5
F/B=0.1					
	5.	3.162	1.2	-14.5	-13.3
	4.	2.512	1.8	-13.0	-11.2
	3.	1.995	2.0	-11.3	-9.3
	2.	1.585	2.1	-9.7	-7.6
	1.	1.259	2.0	-8.2	-6.2
	0.	1.000	1.9	-6.9	-5.0
	-1.	0.794	1.7	-5.8	-4.0
	-2.	0.631	1.6	-4.9	-3.3
	-3.	0.501	1.5	-4.1	-2.6
	-4.	0.398	1.4	-3.5	-2.1
	-5.	0.316	1.3	-3.0	-1.7
	-6.	0.251	1.2	-2.6	-1.3
	-7.	0.200	1.2	-2.2	-1.1
	-8.	0.158	1.1	-2.0	-0.8
	-9.	0.126	1.1	-1.7	-0.7
	-10.	0.100	1.0	-1.6	-0.5
F/B=0.0					
	5.	3.162	1.6	-16.0	-14.4
	4.	2.512	2.0	-14.0	-12.0
	3.	1.995	2.1	-11.9	-9.8
	2.	1.585	2.0	-10.0	-8.0
	1.	1.259	1.8	-8.3	-6.5
	0.	1.000	1.6	-6.8	-5.2
	-1.	0.794	1.3	-5.6	-4.2
	-2.	0.631	1.1	-4.5	-3.4
	-3.	0.501	1.0	-3.7	-2.7
	-4.	0.398	0.8	-3.0	-2.2
	-5.	0.316	0.7	-2.4	-1.7
	-6.	0.251	0.6	-1.9	-1.4
	-7.	0.200	0.5	-1.6	-1.1
	-8.	0.158	0.4	-1.3	-0.9
	-9.	0.126	0.4	-1.1	-0.7
	-10.	0.100	0.3	-0.9	-0.5

Slot Noise
Rectangular and Gaussian RF/IF Response

Table 5

F/B=0.2

C/N (dB)	C/N (PR)	Diff. (dB)	Rect. Noise (dB)	Gaus. Noise (dB)
20.00	100.00	0.34	-28.31	-28.65
19.00	79.43	0.34	-27.86	-28.20
18.00	63.10	0.34	-26.86	-27.20
17.00	50.12	0.34	-25.86	-26.20
16.00	39.81	0.34	-24.86	-25.20
15.00	31.62	0.34	-23.86	-24.20
14.00	25.12	0.34	-22.86	-23.20
13.00	19.95	0.34	-21.86	-22.20
12.00	15.85	0.34	-20.86	-21.20
11.00	12.59	0.34	-19.86	-20.20
10.00	10.00	0.34	-18.86	-19.19
9.00	7.94	0.32	-17.83	-18.15
8.00	6.31	0.25	-16.70	-16.96
7.00	5.01	0.08	-15.36	-15.44
6.00	3.98	0.20	-13.71	-13.51
5.00	3.16	0.51	-11.78	-11.28

F/B=0.1

C/N (dB)	C/N (PR)	Diff. (dB)	Rect. Noise (dB)	Gaus. Noise (dB)
20.00	100.00	0.07	-34.33	-34.26
19.00	79.43	0.07	-33.88	-33.81
18.00	63.10	0.07	-32.88	-32.81
17.00	50.12	0.07	-31.88	-31.81
16.00	39.81	0.07	-30.88	-30.81
15.00	31.62	0.07	-29.88	-29.81
14.00	25.12	0.07	-28.88	-28.81
13.00	19.95	0.07	-27.88	-27.81
12.00	15.85	0.07	-26.88	-26.81
11.00	12.59	0.07	-25.88	-25.81
10.00	10.00	0.08	-24.86	-24.78
9.00	7.94	0.13	-23.74	-23.62
8.00	6.31	0.30	-22.28	-21.98
7.00	5.01	0.63	-20.16	-19.53
6.00	3.98	0.98	-17.43	-16.45
5.00	3.16	1.22	-14.50	-13.28

F/B=0.05

C/N (dB)	C/N (PR)	Diff. (dB)	Rect. Noise (dB)	Gaus. Noise (dB)
20.00	100.00	0.17	-40.35	-40.18
19.00	79.43	0.17	-39.90	-39.73
18.00	63.10	0.17	-38.90	-38.73
17.00	50.12	0.17	-37.90	-37.73
16.00	39.81	0.17	-36.90	-36.73
15.00	31.62	0.17	-35.90	-35.73
14.00	25.12	0.17	-34.90	-34.73
13.00	19.95	0.17	-33.90	-33.73
12.00	15.85	0.17	-32.90	-32.73
11.00	12.59	0.17	-31.89	-31.72
10.00	10.00	0.20	-30.82	-30.62
9.00	7.94	0.36	-29.37	-29.01
8.00	6.31	0.77	-26.87	-26.11
7.00	5.01	1.17	-23.21	-22.04
6.00	3.98	1.40	-19.23	-17.83
5.00	3.16	1.50	-15.56	-14.07

Slot Noise
Rectangular and Gaussian RF/IF Response

Table 5
(cont.)

F/B=0.025

C/N (dB)	C/N (PR)	Diff. (dB)	Rect. Noise (dB)	Gaus. Noise (dB)
20.00	100.00	0.20	-46.37	-46.17
19.00	79.43	0.20	-45.92	-45.73
18.00	63.10	0.20	-44.92	-44.73
17.00	50.12	0.20	-43.92	-43.73
16.00	39.81	0.20	-42.92	-42.73
15.00	31.62	0.20	-41.92	-41.73
14.00	25.12	0.20	-40.92	-40.73
13.00	19.95	0.20	-39.92	-39.73
12.00	15.85	0.20	-38.92	-38.72
11.00	12.59	0.21	-37.89	-37.69
10.00	10.00	0.31	-36.61	-36.30
9.00	7.94	0.73	-34.10	-33.37
8.00	6.31	1.24	-29.63	-28.39
7.00	5.01	1.47	-24.49	-23.02
6.00	3.98	1.55	-19.82	-18.27
5.00	3.16	1.58	-15.88	-14.30

F/B=0.01

C/N (dB)	C/N (PR)	Diff. (dB)	Rect. Noise (dB)	Gaus. Noise (dB)
20.00	100.00	0.20	-54.33	-54.12
19.00	79.43	0.20	-53.88	-53.68
18.00	63.10	0.20	-52.88	-52.68
17.00	50.12	0.20	-51.88	-51.68
16.00	39.81	0.20	-50.88	-50.68
15.00	31.62	0.20	-49.88	-49.68
14.00	25.12	0.20	-48.88	-48.68
13.00	19.95	0.20	-47.88	-47.68
12.00	15.85	0.21	-46.87	-46.67
11.00	12.59	0.27	-45.71	-45.45
10.00	10.00	0.70	-43.21	-42.51
9.00	7.94	1.32	-37.59	-36.27
8.00	6.31	1.53	-30.87	-29.34
7.00	5.01	1.59	-24.93	-23.34
6.00	3.98	1.60	-20.01	-18.41
5.00	3.16	1.61	-15.97	-14.36

F/B=0.005

C/N (dB)	C/N (PR)	Diff. (dB)	Rect. Noise (dB)	Gaus. Noise (dB)
20.00	100.00	0.20	-60.35	-60.14
19.00	79.43	0.20	-59.90	-59.70
18.00	63.10	0.20	-58.90	-58.70
17.00	50.12	0.20	-57.90	-57.70
16.00	39.81	0.20	-56.90	-56.70
15.00	31.62	0.20	-55.90	-55.70
14.00	25.12	0.20	-54.90	-54.70
13.00	19.95	0.20	-53.90	-53.70
12.00	15.85	0.22	-52.87	-52.66
11.00	12.59	0.42	-51.27	-50.84
10.00	10.00	1.17	-46.32	-45.15
9.00	7.94	1.52	-38.43	-36.90
8.00	6.31	1.59	-31.08	-29.49
7.00	5.01	1.60	-25.00	-23.39
6.00	3.98	1.61	-20.03	-18.43
5.00	3.16	1.61	-15.98	-14.37

Slot Noise
Rectangular and Gaussian RF/IF Response

Table 5
(cont.)

F/B=0.0025 C/N (dB)	C/N (PR)	Diff. (dB)	Rect. Noise (dB)	Gaus. Noise (dB)
20.00	100.00	0.20	-66.37	-66.16
19.00	79.43	0.20	-65.92	-65.72
18.00	63.10	0.20	-64.92	-64.72
17.00	50.12	0.20	-63.92	-63.72
16.00	39.81	0.20	-62.92	-62.72
15.00	31.62	0.20	-61.92	-61.72
14.00	25.12	0.20	-60.92	-60.72
13.00	19.95	0.21	-59.92	-59.71
12.00	15.85	0.25	-58.81	-58.56
11.00	12.59	0.80	-55.30	-55.00
10.00	10.00	1.47	-47.64	-46.17
9.00	7.94	1.59	-38.67	-37.08
8.00	6.31	1.60	-31.14	-29.53
7.00	5.01	1.61	-25.01	-23.40
6.00	3.98	1.61	-20.04	-18.43
5.00	3.16	1.61	-15.98	-14.38

F/B=0.001

20.00	100.00	0.20	-74.33	-74.12
19.00	79.43	0.20	-73.88	-73.68
18.00	63.10	0.20	-72.88	-72.68
17.00	50.12	0.20	-71.88	-71.68
16.00	39.81	0.20	-70.88	-70.68
15.00	31.62	0.20	-69.88	-69.68
14.00	25.12	0.21	-68.88	-68.68
13.00	19.95	0.21	-67.87	-67.66
12.00	15.85	0.44	-66.20	-65.76
11.00	12.59	1.36	-58.94	-57.58
10.00	10.00	1.58	-48.09	-46.51
9.00	7.94	1.61	-38.74	-37.13
8.00	6.31	1.61	-31.15	-29.54
7.00	5.01	1.61	-25.02	-23.41
6.00	3.98	1.61	-20.04	-18.43
5.00	3.16	1.61	-15.[illegible]	-14.38

F/B=0.0005

20.00	100.00	0.20	-80.35	-80.14
19.00	79.43	0.20	-79.90	-79.70
18.00	63.10	0.20	-78.90	-78.70
17.00	50.12	0.20	-77.90	-77.70
16.00	39.81	0.20	-76.90	76.70
15.00	31.62	0.20	-75.90	-75.70
14.00	25.12	0.21	-74.90	-74.70
13.00	19.95	0.23	-73.85	-73.62
12.00	15.85	0.83	-70.65	-69.82
11.00	12.59	1.54	-59.66	-58.12
10.00	10.00	1.60	-48.16	-46.55
9.00	7.94	1.61	-38.75	-37.14
8.00	6.31	1.61	-31.15	-29.54
7.00	5.01	1.61	-25.02	-23.41
6.00	3.98	1.61	-20.04	-18.43
5.00	3.16	1.61	-15.99	-14.38

Slot Noise
Rectangular and Gaussian RF/IF Response

Table 5
(cont.)

F/B=0.00025

C/N (dB)	C/N (PR)	Diff. (dB)	Rect. Noise (dB)	Gaus. Noise (dB)
20.00	100.00	0.20	-86.37	-86.16
19.00	79.43	0.20	-85.92	-85.72
18.00	63.10	0.20	-84.92	-84.72
17.00	50.12	0.20	-83.92	-83.72
16.00	39.81	0.20	-82.92	-82.72
15.00	31.62	0.20	-81.92	-81.72
14.00	25.12	0.21	-80.92	-80.72
13.00	19.95	0.28	-79.71	-79.43
12.00	15.85	1.27	-73.22	-71.95
11.00	12.59	1.59	-59.86	-58.27
10.00	10.00	1.61	-48.18	-46.57
9.00	7.94	1.61	-38.75	-37.14
8.00	6.31	1.61	-31.15	-29.54
7.00	5.01	1.61	-25.02	-23.41
6.00	3.98	1.61	-20.04	-18.43
5.00	3.16	1.61	-15.99	-14.38

F/B=0.0001

20.00	100.00	0.20	-94.33	-94.12
19.00	79.43	0.20	-93.88	-93.68
18.00	63.10	0.20	-92.88	-92.68
17.00	50.12	0.20	-91.88	-91.68
16.00	39.81	0.20	-90.88	-90.68
15.00	31.62	0.21	-89.88	-89.68
14.00	25.12	0.21	-88.87	-88.66
13.00	19.95	0.58	-86.69	-86.11
12.00	15.85	1.54	-74.33	-72.79
11.00	12.59	1.61	-59.92	-58.31
10.00	10.00	1.61	-48.18	-46.57
9.00	7.94	1.61	-38.75	-37.14
8.00	6.31	1.61	-31.15	-29.54
7.00	5.01	1.61	-25.02	-23.41
6.00	3.98	1.61	-20.04	-18.43
5.00	3.16	1.61	-15.99	-14.38

F/B=0.0

20.00	100.00	1.61	-358.50	-356.89
19.00	79.43	1.61	-354.22	-352.61
18.00	63.10	1.61	-282.77	-281.16
17.00	50.12	1.61	-225.92	-224.31
16.00	39.81	1.61	-180.65	-179.04
15.00	31.62	1.61	-144.59	-142.98
14.00	25.12	1.61	-115.84	-114.23
13.00	19.95	1.61	-92.91	-91.30
12.00	15.85	1.61	-74.58	-72.97
11.00	12.59	1.61	-59.93	-58.32
10.00	10.00	1.61	-48.18	-46.57
9.00	7.94	1.61	-38.75	-37.14
8.00	6.31	1.61	-31.15	-29.54
7.00	5.01	1.61	-25.02	-23.41
6.00	3.98	1.61	-20.04	-18.43
5.00	3.16	1.61	-15.99	-14.38

Slot Noise
Rectangular and Gaussian RF/IF Response

Table 5
(cont.)

	C/N (dB)	C/N (PR)	Diff. (dB)	Avg. Noise (dB)
F/B=0.2				
	5.	3.162	0.3	-11.4
	4.	2.512	0.6	-10.2
	3.	1.995	0.8	-8.8
	2.	1.585	0.9	-7.6
	1.	1.259	0.9	-6.4
	0.	1.000	1.0	-5.4
	-1.	0.794	1.0	-4.6
	-2.	0.631	1.0	-3.9
	-3.	0.501	0.9	-3.3
	-4.	0.398	0.9	-2.8
	-5.	0.316	0.9	-2.4
	-6.	0.251	0.9	-2.1
	-7.	0.200	0.9	-1.9
	-8.	0.158	0.9	-1.7
	-9.	0.126	0.9	-1.5
	-10.	0.100	0.9	-1.4
F/B=0.1				
	5.	3.162	0.6	-13.9
	4.	2.512	0.9	-12.1
	3.	1.995	1.0	-10.3
	2.	1.585	1.0	-8.6
	1.	1.259	1.0	-7.2
	0.	1.000	0.9	-6.0
	-1.	0.794	0.9	-4.9
	-2.	0.631	0.8	-4.1
	-3.	0.501	0.8	-3.3
	-4.	0.398	0.7	-2.8
	-5.	0.316	0.7	-2.3
	-6.	0.251	0.6	-1.9
	-7.	0.200	0.6	-1.6
	-8.	0.158	0.6	-1.4
	-9.	0.126	0.5	-1.2
	-10.	0.100	0.5	-1.0
F/B=0.0				
	5.	3.162	0.8	-15.2
	4.	2.512	1.0	-13.0
	3.	1.995	1.0	-10.9
	2.	1.585	1.0	-9.0
	1.	1.259	0.9	-7.4
	0.	1.000	0.8	-6.0
	-1.	0.794	0.7	-4.9
	-2.	0.631	0.6	-3.9
	-3.	0.501	0.5	-3.2
	-4.	0.398	0.4	-2.6
	-5.	0.316	0.3	-2.1
	-6.	0.251	0.3	-1.7
	-7.	0.200	0.3	-1.3
	-8.	0.158	0.2	-1.1
	-9.	0.126	0.2	-0.9
	-10.	0.100	0.2	-0.7

Averaged Slot Noise

Table 6

F/B=0.2	C/N (dB)	C/N (PR)	Diff. (dB)	Avg. Noise (dB)
	20.00	100.00	0.27	-29.03
	19.00	79.43	0.27	-28.03
	18.00	63.10	0.27	-27.03
	17.00	50.12	0.27	-26.03
	16.00	39.81	0.27	-25.03
	15.00	31.62	0.27	-24.03
	14.00	25.12	0.27	-23.03
	13.00	19.95	0.27	-22.03
	12.00	15.85	0.27	-21.03
	11.00	12.59	0.27	-20.03
	10.00	10.00	0.27	-19.02
	9.00	7.94	0.26	-17.99
	8.00	6.31	0.23	-16.83
	7.00	5.01	0.14	-15.40
	6.00	3.98	0.00	-13.61
	5.00	3.16	0.15	-11.53
F/B=0.1				
	20.00	100.00	0.07	-34.85
	19.00	79.43	0.07	-33.85
	18.00	63.10	0.07	-32.85
	17.00	50.12	0.07	-31.85
	16.00	39.81	0.07	-30.85
	15.00	31.62	0.07	-29.85
	14.00	25.12	0.07	-28.85
	13.00	19.95	0.07	-27.85
	12.00	15.85	0.07	-26.85
	11.00	12.59	0.07	-25.84
	10.00	10.00	0.06	-24.82
	9.00	7.94	0.04	-23.68
	8.00	6.31	0.05	-22.13
	7.00	5.01	0.21	-19.84
	6.00	3.98	0.39	-16.94
	5.00	3.16	0.51	-13.89
F/B=0.05				
	20.00	100.00	0.02	-40.82
	19.00	79.43	0.02	-39.82
	18.00	63.10	0.02	-38.82
	17.00	50.12	0.02	-37.82
	16.00	39.81	0.02	-36.82
	15.00	31.62	0.02	-35.82
	14.00	25.12	0.02	-34.82
	13.00	19.95	0.02	-33.82
	12.00	15.85	0.02	-32.82
	11.00	12.59	0.02	-31.81
	10.00	10.00	0.00	-30.72
	9.00	7.94	0.08	-29.19
	8.00	6.31	0.28	-26.49
	7.00	5.01	0.48	-22.63
	6.00	3.98	0.60	-18.53
	5.00	3.16	0.65	-14.82

Averaged Slot Noise

Table 6
(cont.)

F/B=0.025

C/N (dB)	C/N (PR)	Diff. (dB)	Avg. Noise (dB)
20.00	100.00	0.00	-46.82
19.00	79.43	0.00	-45.82
18.00	63.10	0.00	-44.82
17.00	50.12	0.00	-43.82
16.00	39.81	0.00	-42.82
15.00	31.62	0.00	-41.82
14.00	25.12	0.00	-40.82
13.00	19.95	0.00	-39.82
12.00	15.85	0.00	-38.82
11.00	12.59	0.00	-37.79
10.00	10.00	0.05	-36.45
9.00	7.94	0.26	-33.73
8.00	6.31	0.52	-29.01
7.00	5.01	0.63	-23.76
6.00	3.98	0.67	-19.05
5.00	3.16	0.69	-15.09

F/B=0.01

C/N (dB)	C/N (PR)	Diff. (dB)	Avg. Noise (dB)
20.00	100.00	0.00	-54.78
19.00	79.43	0.00	-53.78
18.00	63.10	0.00	-52.78
17.00	50.12	0.00	-51.78
16.00	39.81	0.00	-50.78
15.00	31.62	0.00	-49.78
14.00	25.12	0.00	-48.78
13.00	19.95	0.00	-47.78
12.00	15.85	0.00	-46.77
11.00	12.59	0.03	-45.58
10.00	10.00	0.25	-42.86
9.00	7.94	0.56	-36.93
8.00	6.31	0.66	-30.11
7.00	5.01	0.69	-24.14
6.00	3.98	0.70	-19.21
5.00	3.16	0.70	-15.17

F/B=0.005

C/N (dB)	C/N (PR)	Diff. (dB)	Avg. Noise (dB)
20.00	100.00	0.00	-60.80
19.00	79.43	0.00	-59.80
18.00	63.10	0.00	-58.80
17.00	50.12	0.00	-57.80
16.00	39.81	0.00	-56.80
15.00	31.62	0.00	-55.80
14.00	25.12	0.00	-54.80
13.00	19.95	0.00	-53.80
12.00	15.85	0.01	-52.76
11.00	12.59	0.11	-51.05
10.00	10.00	0.48	-45.74
9.00	7.94	0.66	-37.67
8.00	6.31	0.69	-30.29
7.00	5.01	0.70	-24.19
6.00	3.98	0.70	-19.23
5.00	3.16	0.70	-15.18

Averaged Slot Noise

Table 6
(cont.)

F/B=0.0025	C/N (dB)	C/N (PR)	Diff. (dB)	Avg. Noise (dB)
	20.00	100.00	0.00	-66.82
	19.00	79.43	0.00	-65.82
	18.00	63.10	0.00	-64.82
	17.00	50.12	0.00	-63.82
	16.00	39.81	0.00	-62.82
	15.00	31.62	0.00	-61.82
	14.00	25.12	0.00	-60.82
	13.00	19.95	0.00	-59.82
	12.00	15.85	0.02	-58.68
	11.00	12.59	0.30	-55.40
	10.00	10.00	0.63	-46.90
	9.00	7.94	0.69	-37.87
	8.00	6.31	0.70	-30.33
	7.00	5.01	0.70	-24.21
	6.00	3.98	0.70	-19.23
	5.00	3.16	0.70	-15.18
F/B=0.001				
	20.00	100.00	0.00	-74.78
	19.00	79.43	0.00	-73.78
	18.00	63.10	0.00	-72.78
	17.00	50.12	0.00	-71.78
	16.00	39.81	0.00	-70.78
	15.00	31.62	0.00	-69.78
	14.00	25.12	0.00	-68.78
	13.00	19.95	0.00	-67.76
	12.00	15.85	0.12	-65.98
	11.00	12.59	0.58	-58.26
	10.00	10.00	0.69	-47.30
	9.00	7.94	0.70	-37.93
	8.00	6.31	0.70	-30.35
	7.00	5.01	0.70	-24.21
	6.00	3.98	0.70	-19.24
	5.00	3.16	0.70	-15.18
F/B=0.0005				
	20.00	100.00	0.00	-80.80
	19.00	79.43	0.00	-79.80
	18.00	63.10	0.00	-78.80
	17.00	50.12	0.00	-77.80
	16.00	39.81	0.00	-76.80
	15.00	31.62	0.00	-75.80
	14.00	25.12	0.00	-74.80
	13.00	19.95	0.01	-73.73
	12.00	15.85	0.31	-70.24
	11.00	12.59	0.67	-58.89
	10.00	10.00	0.70	-47.36
	9.00	7.94	0.70	-37.94
	8.00	6.31	0.70	-30.35
	7.00	5.01	0.70	-24.21
	6.00	3.98	0.70	-19.24
	5.00	3.16	0.70	-15.18

Averaged Slot Noise

Table 6
(cont.)

F/B	C/N (dB)	C/N (PR)	Diff. (dB)	Avg. Noise (dB)
F/B=0.00025				
	20.00	100.00	0.00	-86.82
	19.00	79.43	0.00	-85.82
	18.00	63.10	0.00	-84.82
	17.00	50.12	0.00	-83.82
	16.00	39.81	0.00	-82.82
	15.00	31.62	·0.	-81.82
	14.00	25.12	0.00	-80.82
	13.00	19.95	0.04	-79.57
	12.00	15.85	0.53	-72.58
	11.00	12.59	0.69	-59.06
	10.00	10.00	0.70	-47.37
	9.00	7.94	0.70	-37.94
	8.00	6.31	0.70	-30.35
	7.00	5.01	0.70	-24.21
	6.00	3.98	0.70	-19.24
	5.00	3.16	0.70	-15.18
F/B=0.0001				
	20.00	100.00	0.	-94.78
	19.00	79.43	0.	-93.78
	18.00	63.10	0.	-92.78
	17.00	50.12	0.	-91.78
	16.00	39.81	0.	-90.78
	15.00	31.62	0.00	-89.78
	14.00	25.12	0.00	-88.77
	13.00	19.95	0.19	-86.40
	12.00	15.85	0.67	-73.56
	11.00	12.59	0.70	-59.11
	10.00	10.00	0.70	-47.38
	9.00	7.94	0.70	-37.94
	8.00	6.31	0.70	-30.35
	7.00	5.01	0.70	-24.21
	6.00	3.98	0.70	-19.24
	5.00	3.16	0.70	-15.18
F/B=0.0				
	20.00	100.00	0.70	-357.70
	19.00	79.43	0.70	-353.42
	18.00	63.10	0.70	-281.97
	17.00	50.12	0.70	-225.11
	16.00	39.81	0.70	-179.84
	15.00	31.62	0.70	-143.78
	14.00	25.12	0.70	-115.04
	13.00	19.95	0.70	-92.10
	12.00	15.85	0.70	-73.78
	11.00	12.59	0.70	-59.12
	10.00	10.00	0.70	-47.38
	9.00	7.94	0.70	-37.94
	8.00	6.31	0.70	-30.35
	7.00	5.01	0.70	-24.21
	6.00	3.98	0.70	-19.24
	5.00	3.16	0.70	-15.18

Averaged Slot Noise

Table 6
(cont.)

3.4 The formulas used to generate the slot noise tables are given later in this report. In some cases, there is a slight error associated with the equation approximations. The error limits are noted later. In the tabulated results, F is the baseband frequency of the center of the noise slot; B is the IF bandwidth C/N is carrier to (thermal) noise ratio. PR indicates power ratio. DIFF is the difference between the two slot noise values. This gives a good check on the dependency of the slot noise on IF response. Gaussian and rectangular IF responses form reasonable limiting cases which limit the realistic range of IF response shapes encountered in practice. All slot noise values have been normalized to the value of slot noise at F/B = 0 and C/N = -dB for he respective IF response.

3.5 Following this set of values, the averaged values have been computed. These values are the algebraic average of the values for the Gaussian and the rectangular IF bandwidth responses. The DIFF column indicates the difference between the averaged IF response and either of the Gaussian and rectangular IF responses. This column gives an estimate of the variation to be expected between averaged theoretical quieting curve characteristics and actual quieting curve measurements. The averaged noise values have been normalized to the averaged slot noise for F/B = 0 and C/N = - (dB).

3.6 There are a couple of common FM receiver specifications related to the slot noise performance of the receiver in the A, B, and upper C regions. One of the specifications is 20 dB (and occasionally 30 dB) noise quieting. The 20 dB (or 30 dB) quieting specification is the RSL at which the noise in a specified baseband slot is 20 dB (or 30 dB) lower than the noise in the slot with no RSL present at the receiver input. Theoretical, 20 dB and 30 dB C/N ratios are listed on the next page. The C/N values can be converted to RSL using receiver parameters and the previous formulas.

3.7 The other specification is FM slot noise threshold or simply FM threshold. There is not complete agreement by sources as to the definition of FM threshold. The most common definition, however, is the RSL at which the receiver baseband slot noise has increased 2 dB for a 1 dB reduction in RSL. This is often determined graphically with quieting curves by extending the straight line of region C into region B and defining FM threshold as the RSL at which slot noise is one dB above the straight line extension. Generally, FM threshold is measured in a narrow (nominal 3.1 KHz wide) noise slot. Theoretical narrow slot FM threshold values are listed on the following page. Occasionally, however, it is desirable to measure FM threshold in a relatively wide slot. For example, if the FM receiver baseband contains a wideband video or digital circuit, the FM noise threshold over the entire bandwidth of the baseband signal would be appropriate. The theoretical wide noise slot FM thresholds are tabulated on the next page. For the FM threshold charts, (f/B) is the narrow slot baseband frequency divided by the IF

	C/N(dB)		
f/B	Rect. IF	Gaus. IF	Avg. IF
0.2	12.6	10.9	11.8
0.1	7.2	7.2	7.2
0.05	6.3	6.5	6.4
0.025	6.1	6.4	6.3
0.01	6.0	6.3	6.2
0.001	6.0	6.3	6.2
0.0001	6.0	6.3	6.2
0.0	6.0	6.3	6.2

Theoretical 20 dB Noise Quieting
(narrow slots)

	C/N(dB)		
f/B	Rect. IF	Gaus. IF	Avg. IF
0.2	22.6	20.9	21.8
0.1	15.8	15.2	15.5
0.05	9.6	9.6	9.6
0.025	8.1	8.3	8.2
0.01	7.9	8.1	8.0
0.001	7.8	8.1	8.0
0.0001	7.8	8.1	8.0
0.0	7.8	8.1	8.0

Theoretical 30 dB Noise Quieting
(narrow slots)

Table 7

f/B	Rectangular RF/IF C/N (dB)	Gaussian RF/IF C/N (dB)	Averaged Response C/N (dB)
0.2	6.19	6.69	6.44
0.1	7.56	7.85	7.71
0.05	8.58	8.79	8.69
0.025	9.39	9.56	9.48
0.0158	9.85	10.00	9.93
0.0147	9.92	10.07	10.0
0.0134	10.00	10.15	10.08
0.01	10.26	10.40	10.33
0.005	10.84	10.94	10.89
0.0025	11.31	11.41	11.36
0.001	11.88	11.97	11.93
0.0005	12.26	12.35	12.31
0.00025	12.62	12.70	12.66
0.0001	13.04	13.11	13.08
0.0	Undefined	Undefined	Undefined

Theoretical FM Noise Thresholds

(narrow slots)

Table 8

Table 9

C/N (dB)

$\frac{(f/B)_{min}}{(f/B)_{max}}$	$(f/B)_{max}=0.2$ Rect. IF	$(f/B)_{max}=0.2$ Gaus. IF	$(f/B)_{max}=0.1$ Rect. IF	$(f/B)_{max}=0.1$ Gaus. IF	$(f/B)_{max}=0.01$ Rect. IF	$(f/B)_{max}=0.01$ Gaus. IF	$(f/B)_{max}=0.001$ Rect. IF	$(f/B)_{max}=0.001$ Gaus. IF	$(f/B)_{max}=0.0001$ Rect. IF	$(f/B)_{max}=0.0001$ Gaus. IF
0.999	6.2	6.7	7.6	7.9	10.3	10.4	11.9	12.0	13.0	13.1
0.9	6.3	6.8	7.6	7.9	10.3	10.4	11.9	12.0	13.1	13.1
0.5	6.8	7.2	8.0	8.2	10.5	10.6	12.0	12.1	13.2	13.2
0.1	7.2	7.6	8.3	8.5	10.7	10.8	12.2	12.2	13.3	13.3
0.0	7.3	7.7	8.4	8.6	10.7	10.8	12.2	12.3	13.3	13.3

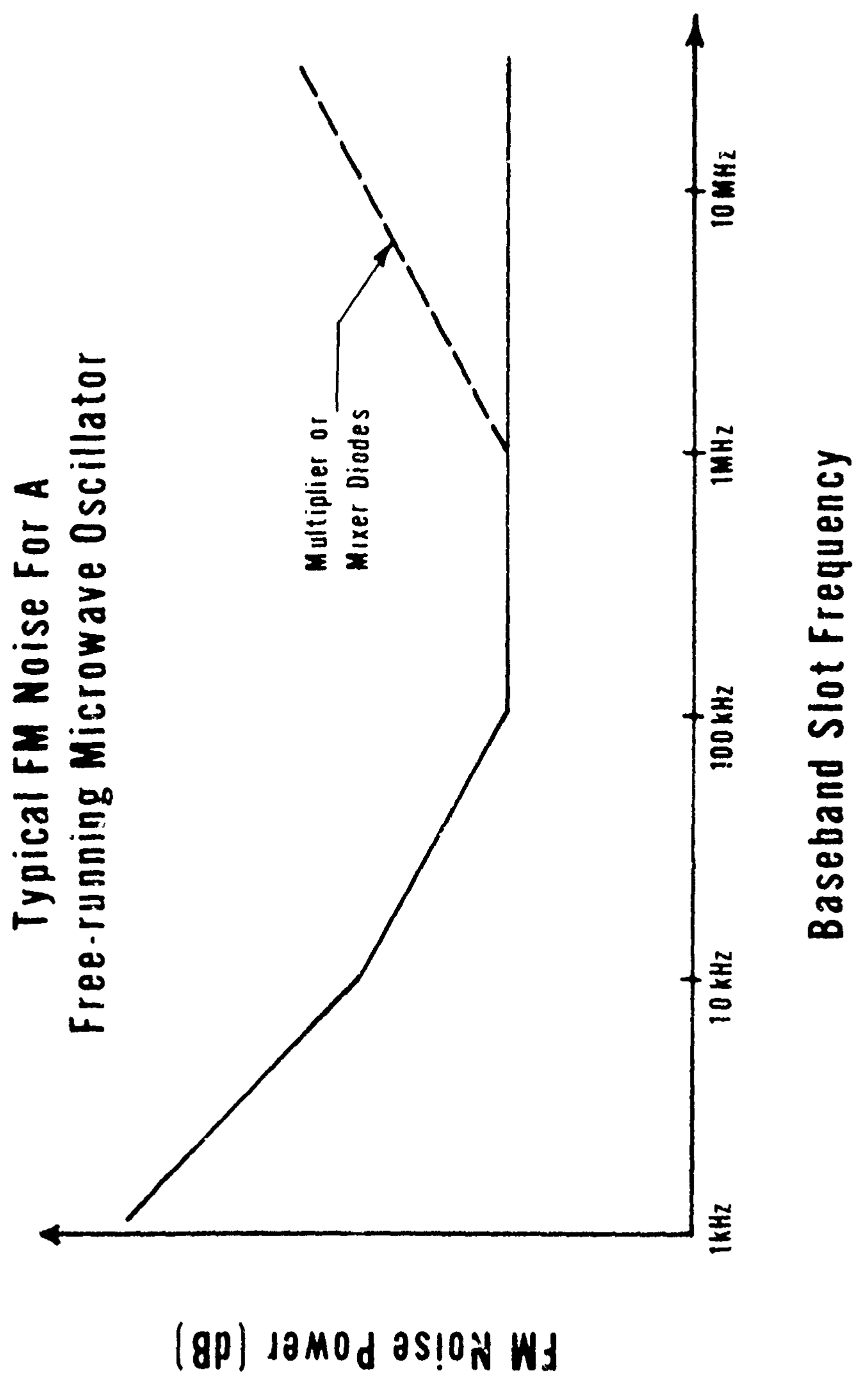
Theoretical FM Noise Thresholds

(wide noise slots)

bandwidth, (f/B) max is the highest baseband slot frequency divided by IF bandwidth, and (f/B) min is the lowest baseband slot frequency divided by IF bandwidth. Note that a wide slot FM threshold always occurs within 1 dB of the FM threshold for a narrow slot which has baseband frequency equal to the maximum frequency of the wide slot.

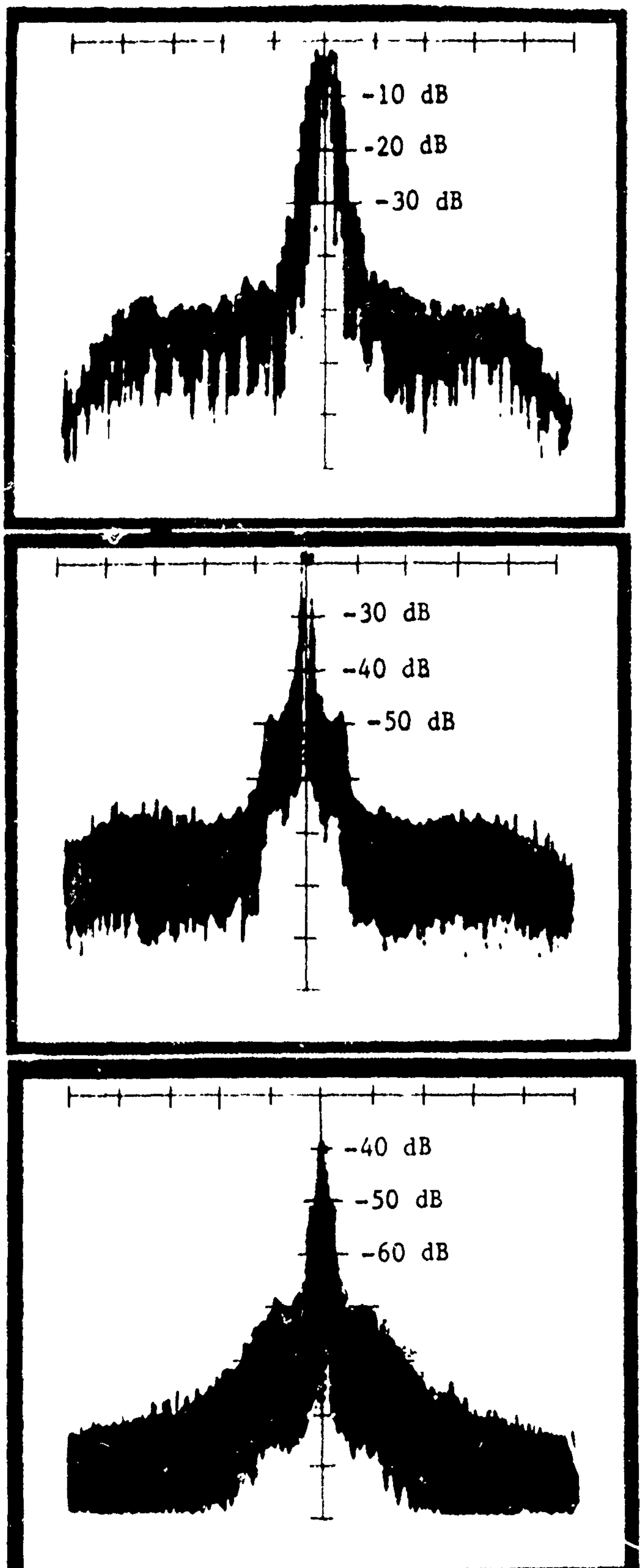
3.8 The last region of interest is D. By conventional linearized methods, (e.g., see Rowe and Panter) it can be shown that in the D region, the noise is the normal f^2 (6 dB per octave) thermal noise plus the time derivative of any phase noise associated with the received carrier. The differentiated phase noise is called frequency noise. It is distinctly different noise than the f^2 receiver thermal noise. For information regarding phase and frequency noise, see references listed in bibliography. In the D region, the f^2 noise is negligible and the dominate noise source is the frequency noise of the received carrier. At a sufficiently high, stable carrier to thermal noise ratio with no multipath and the carrier tuned to the center of a symetric IF, the only significant frequency noise sources are the various oscillators in the transmitter and receiver.

3.9 The frequency noise of an oscillator is a complex phenomenon. In general, the phase noise of an oscillator varies in power roughly as $1 / f^2$ where f is the frequency of the measurement slot away from the carrier frequency. The noise at the baseband of an FM demodulator, however, is the time derivative of the phase noise (for high carrier to noise ratios). Since the power response of a differentiator is proportional to f^2, the frequency noise appearing at the baseband of the radio is essentially flat (independent of baseband frequency). After review of information regarding fundamental frequency (unmultiplied) free running microwave oscillators from various vendors, it appears that this is, in fact, the case for baseband frequencies greater than roughly 100 kHz. Below this frequency, the noise gradually increases. The frequency noise within the first few kilohertz of the baseband can be several times the noise in the mid region of the baseband. As Baprawski et al mention, use of properly optimized phase-locked loops can be used to reduce the noise near the low end of the baseband. Due to feedback loop frequency response requirements, present phase-locked frequency sources can only be used to reduce noise in roughly the first one half megahertz of the radio baseband. Often the output of an oscillator is multiplied to derive a higher frequency. As Payne points out, multiplying the output of the oscillator increases the effective deviation of the frequency noise. Specifically, noise increase is 20 log M where M is composite oscillator multiplication. For example, doubling the oscillator frequency causes a 6 dB increase in frequency noise due to doubling the effective deviation of the frequency noise components. Payne also mentions that in the gigahertz region, the noise figure of the multiplier diodes can be a significant factor. Up or down converter mixer diodes can have the same effect. The noise figure of the diodes can cause a significant increase in slot noise above roughly one megahertz in the radio baseband. The following page graphs typical FM receiver baseband slot noise due to a free running unmodulated M/W oscillator.



Typical Microwave Oscillator Noise

Figure 5



Horizontal: 1 kHz/Div.

Measurement Bandwidth:
100 Hz

Horizontal: 5 kHz/Div.

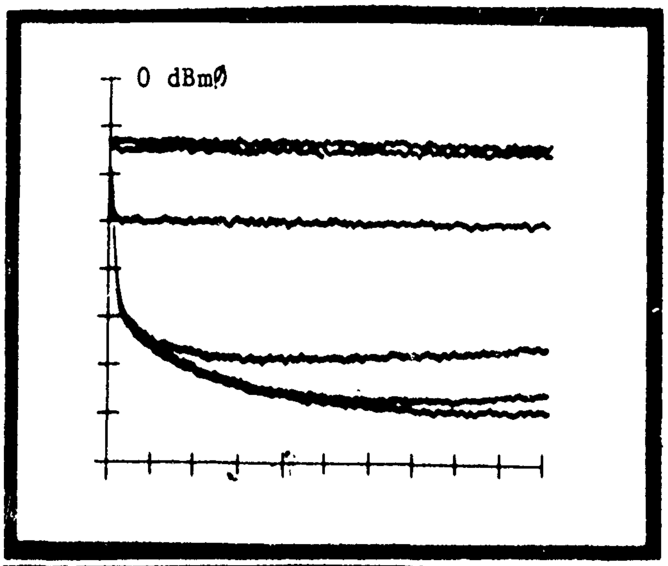
Measurement Bandwidth:
100 Hz

Horizontal: 20 kHz/Div.

Measurement Bandwidth:
300 Hz

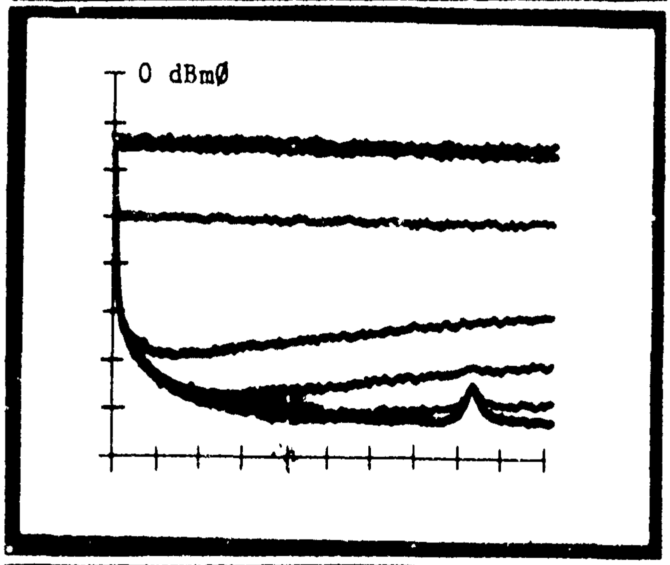
TROPO M/W Transmitter
VCO Spectrum

Figure 6



Vertical: 10 dB/Div.
Horizontal: 20 kHz/Div.
(0 to 200 kHz)

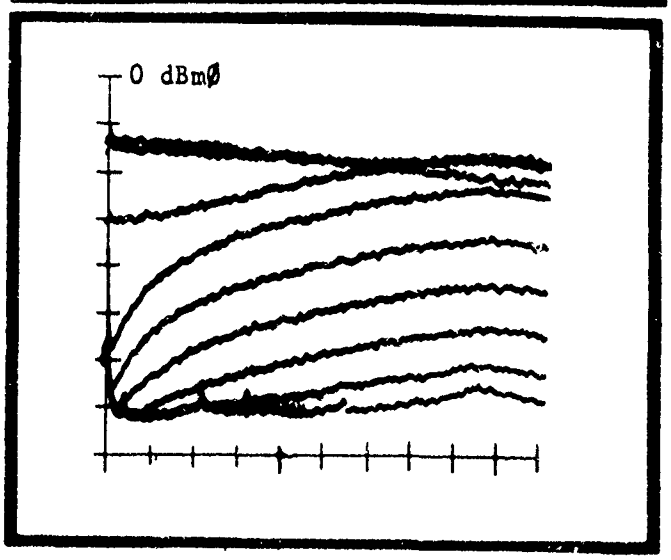
Note: This LOS M/W receiver is not the one used in the rest of this report.



Vertical: 10 dB/Div.
Horizontal: 50 kHz/Div.
(0 to 500 kHz)

RSLs(dBm) = -100, -90, -80, -70, -60, -50, -40, -30, -20

Measurement Bandwidth: 3 kHz



Vertical: 10 dB/Div.
Horizontal: 1 MHz/Div.
(0 to 10 MHz)

LOS M/W Receiver
no de-emphasis

Baseband Noise Spectrum
for Various RSLs

Figure 7

3.10 There are several oscillators in a microwave system. The primary oscillators of interest are the local oscillators for up and down frequency conversion and the frequency modulated oscillator at the transmitter (voltage controlled oscillator-VCO). The up and down converter local oscillators can be phase-locked to reduce low frequency noise. As Lawson and Uhlenbeck have noted, certain hybrid configurations allow the noise sidebands of the oscillator to be suppressed. These methods can not be used on the modulated oscillator, however, since any method used to suppress the inherent noise of the oscillator will also suppress the modulation. Therefore, the oscillator to be modulated <u>must</u> have inherently low noise. If the output of this oscillator is multiplied, a much quieter oscillator is required. It should be mentioned that the transmit oscillators are usually frequency stablized by phase locked loops. The loop bandwidths of these systems are quite narrow (often less than a Hertz). These loops will not suppress noise in the radio baseband greater than a few Hertz above zero frequency. It can be anticipated that the baseband noise of a voltage controlled M/W oscillator used for FM will be similar to the noise of a free running M/W oscillator.

3.11 To point out the complex nature of oscillator noise, spectrum noise plots of the sideband noise spectrum of a microwave voltage controlled oscillator (with Automatic Frequency Control circuitry disabled) were taken. The plots are on the next page. The sideband frequency distribution is a combination of both amplitude modulation noise (in phase noise) as well as phase/frequency noise (quadrature noise). Therefore, a direct relation between the frequency spectrum and noise in the baseband is not possible. However, the noise process is obviously complex in the near carrier region.

3.12 To illustrate the thermal and phase noise in a typical microwave receiver, the baseband noise spectrum was plotted for various received signal levels using a spectrum analyzer. The plots are on the next page. Note the oscillator noise at the low baseband frequencies at high C/Ns and the crossing of the noise slots for low C/Ns.

3.13 Having determined the basic slot noise as a function of RSL, the next item of immediate interest is the performance of the baseband signal versus RSL. As long as the received signal is hard limited by the limiter prior to demodulation and as long as the carrier to noise ratio (C/N) is large, the received signal at the baseband will vary directly as the transmitted baseband signal and will be completely independent of RSL. Part of an FM demodulator is an envelope detector. All incoherent detectors, including envelope detectors, exhibit a baseband signal threshold. That is to say, for low C/Ns, the baseband signal will be suppressed. Several people (e.g., Middleton, Rice, and Stumpers) have derived equations relating FM demodulator input modulated signal to noise ratio and output baseband signal suppression for a sine wave baseband signal. Schwartz shows that the same equation for baseband

C/N (dB)	C/N (PR)	Signal Level (dB)
10.	10.0000	0.0
9.	7.9433	0.0
8.	6.3096	0.0
7.	5.0119	-0.1
6.	3.9811	-0.2
5.	3.1623	-0.4
4.	2.5119	-0.7
3.	1.9953	-1.3
2.	1.5849	-2.0
1.	1.2589	-2.9
0.	1.0000	-4.0
-1.	0.7943	-5.2
-2.	0.6310	-6.6
-3.	0.5012	-8.1
-4.	0.3981	-9.7
-5.	0.3162	-11.3
-6.	0.2512	-13.1
-7.	0.1995	-14.9
-8.	0.1585	-16.7
-9.	0.1259	-18.5
-10.	0.1000	-20.4
-11.	0.0794	-22.3
-12.	0.0631	-24.3
-13.	0.0501	-26.2
-14.	0.0398	-28.2
-15.	0.0316	-30.1
-16.	0.0251	-32.1
-17.	0.0200	-34.1
-18.	0.0158	-36.1
-19.	0.0126	-38.1
-20.	0.0100	-40.0
-21.	0.0079	-42.0
-22.	0.0063	-44.0
-23.	0.0050	-46.0
-24.	0.0040	-48.0
-25.	0.0032	-50.0
-26.	0.0025	-52.0
-27.	0.0020	-54.0
-28.	0.0016	-56.0
-29.	0.0013	-58.0
-30.	0.0010	-60.0
-31.	0.0008	-62.0
-32.	0.0006	-64.0
-33.	0.0005	-66.0
-34.	0.0004	-68.0
-35.	0.0003	-70.0
-36.	0.0003	-72.0
-37.	0.0002	-74.0
-38.	0.0002	-76.0
-39.	0.0001	-78.0
-40.	0.0001	-80.0

Baseband Signal Suppression

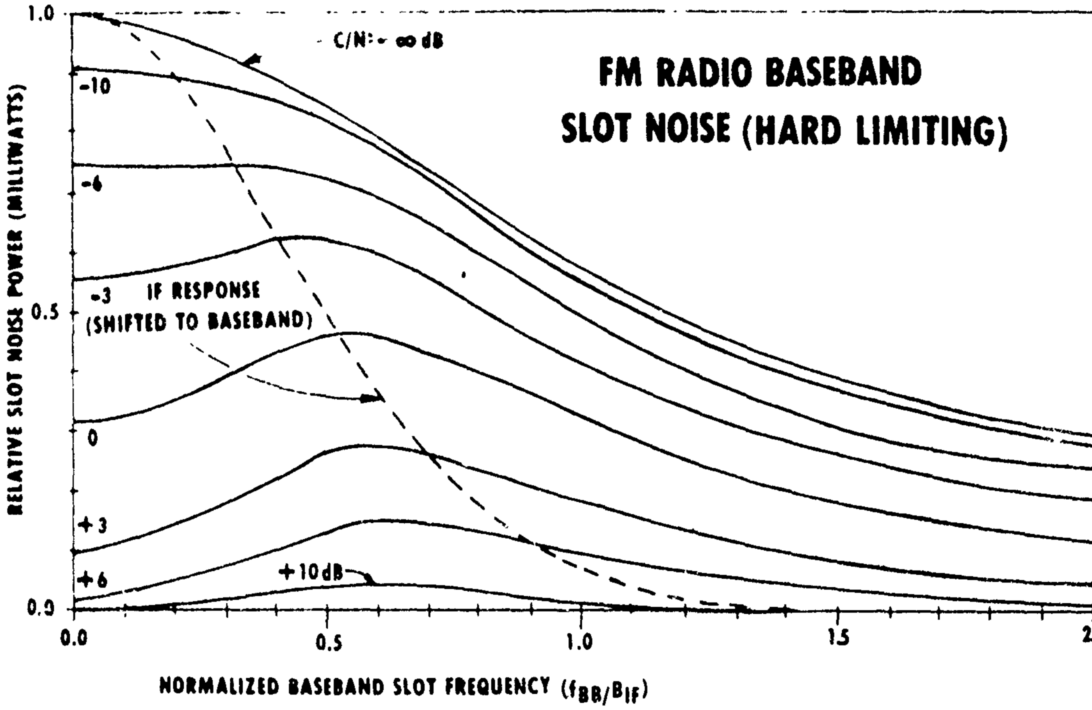
Table 10

signal suppression holds for any arbitrary baseband signal. The equation for baseband signal suppression will be given later in this report. The relation between baseband signal suppression and receiver input carrier to noise ratio is listed on the next page. The dB factor listed is (algebraicly) added to the normal baseband dBm or dBmØ power level. Receiver carrier to noise ratio can be converted to received signal level (RSL) based on FM receiver parameters using formulas listed later in this report.

3.14 One of the assumptions so far has been that the microwave receiver was hard limiting the received signal prior to demodulation. This assumption requires us to believe that all microwave receiver limiters are driven to saturation with just the thermal noise from the front end of the receiver. This is a highly questionable assumption. Loss of hard limiting effects both the baseband signal and noise properties of an FM receiver.

3.15 Like the problem of FM noise for low C/N, the problem of demodulated FM signals and noise with arbitrary or no limiting is complex. Middleton has solved this problem for an FM receiver with Gaussian IF frequency response. His results indicate that as limiter action transitious from hard limiting to soft (partial) limiting to no limiting, the slot noise with no carrier present transitions to less and less noise. Also, the baseband noise spectrum changes dramatically as limiting is lost. Middleton's results for the two extreme cases of limiting were used to determine the following normalized baseband noise spectrum charts. In absolute power the noise associated with "1.0 milliwatt" on the "hard limiting" curve is greater than the absolute power associated with "1.0 milliwatt" on the "no limiting" chart. The most obvious differences between the two curves is the change in shape. This is interesting, but the curves also indicate a more important consideration. For a given slot frequency and given C/N greater than 0 dB, the slot noise is much greater with no limiting than it is with full limiting. We can anticipate that if a receiver limiter action starts going soft for low C/Ns, the slot noise will increase more than usual. At very low C/Ns however, we can expect slot noise to be lower than it would be with hard limiting. The slot noise in the baseband of a M/W receiver varies depending on the amount of limiter action. The baseband noise spectrum for an FM receiver with rectangular IF and no limiting has not, to the author's knowledge, been derived. However, Stumpers and Wang have derived the baseband noise spectrum for a rectangular IF FM receiver with hard limiting. Their results indicate that, when compared with the Gaussian IF results for low carrier to noise ratios, baseband noise falls off more quickly with increasing baseband frequency and, for the intermediate carrier to noise ratios, the baseband noise peaks occur at lower baseband frequencies. The overall shapes of the baseband noise spectra are similar.

3.16 Using the results of Middleton, Rice and Stumper's, the slot noise for hard limiting and no limiting were derived. These results are summarized on the next page. These values are derived from truncated infinite


FM RADIO BASEBAND
SLOT NOISE (HARD LIMITING)
C/N: - ∞ dB
-10
-6
-3
IF RESPONSE
(SHIFTED TO BASEBAND)
0
+3
+6
+10dB
RELATIVE SLOT NOISE POWER (MILLIWATTS)
1.0
0.5
0.0
0.0
0.5
1.0
1.5
2.0
NORMALIZED BASEBAND SLOT FREQUENCY (f_{BB}/B_{IF})

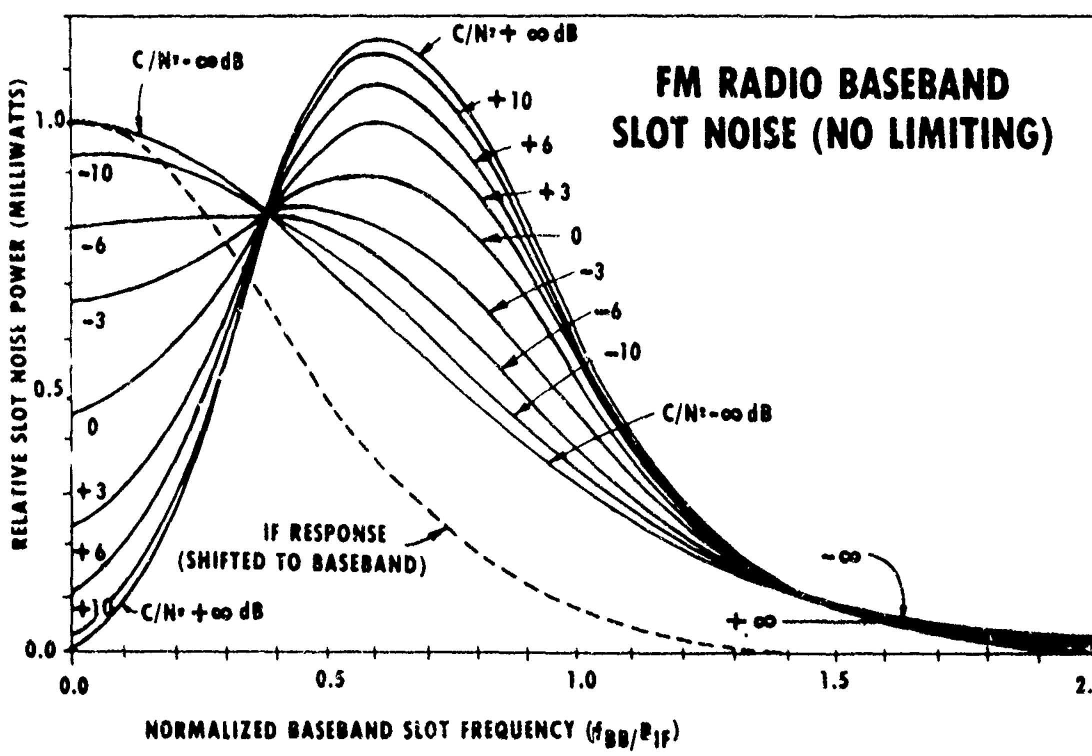

FM RADIO BASEBAND
SLOT NOISE (NO LIMITING)
C/N: + ∞ dB
+10
+6
+3
0
-3
-6
-10
C/N: -∞ dB
C/N: -∞dB
-10
-6
-3
0
+3
+6
+10
C/N: +∞ dB
IF RESPONSE
(SHIFTED TO BASEBAND)
-∞
+∞
RELATIVE SLOT NOISE POWER (MILLIWATTS)
1.0
0.5
0.0
0.0
0.5
1.0
1.5
2.0
NORMALIZED BASEBAND SLOT FREQUENCY (f_{BB}/B_{IF})

Figure 8

Table 11

f/B	Rectangular Response Hard Limiting	Rectangular Response No Limiting	Gaussian Response Hard Limiting	Gaussian Response No Limiting	Averaged Response Hard Limiting	Averaged Response No Limiting
0.000	0.00*	0.00**	0.00*	0.00**	0.00	0.00
0.025	-0.18	-0.30	-0.01	0.00	-0.10	-0.15
0.050	-0.36	-0.61	-0.01	-0.02	-0.19	-0.32
0.075	-0.53	-0.93	-0.02	-0.04	-0.28	-0.49
0.100	-0.71	-1.26	-0.04	-0.06	-0.38	-0.66
0.125	-0.89	-1.59	-0.06	-0.10	-0.48	-0.85
0.150	-1.07	-1.93	-0.08	-0.14	-0.58	-1.04
0.175	-1.25	-2.28	-0.11	-0.19	-0.68	-1.24
0.200	-1.43	-2.63	-0.14	-0.25	-0.79	-1.44
0.250	-1.79	-3.38	-0.22	-0.39	-1.01	-1.89
0.300	-2.15	-4.16	-0.31	-0.56	-1.23	-2.36
0.350	-2.50	-4.98	-0.42	-0.76	-1.46	-2.87
0.400	-2.86	-5.85	-0.54	-1.00	-1.70	-3.43
0.450	-3.21	-6.77	-0.68	-1.26	-1.95	-4.02
0.500	-3.54	-7.75	-0.83	-1.55	-2.19	-4.65

Results have been normalized to the C/N = $-\infty$, f/B = 0.00 value.

*In absolute level, the rectangular response C/N = $-\infty$,
f/B = 0.00 value is 0.21 dB greater than the corresponding Gaussian value.

**In absolute level, the rectangular response C/N = $-\infty$,
f/B = 0.00 value is 1.25 dB greater than the corresponding Gaussian value.

Slot Noise (dB) Vs Baseband Frequency
(no carrier present)

series summation of terms related to the IF characteristic. It is interesting that for the hard limiting case, the series converged very slowly. The values given represent 250,000 terms in the series. This many terms were needed to get the answer accurate to within a couple of units in the fourth place. These results agree well with the calculated values of Rice and Stumpers. Although the attempt was made, the slot noise distribution for no limiting and no carrier present could not be verified experimentally.

3.17 Limiter action also affects the demodulated baseband signal. It appears that the solution for an arbitrary baseband signal in the presence of arbitrary limiting has not been solved. Middleton has determined the effect of soft limiting on the suppression of a baseband test tone. His results indicate that as limiter action goes soft, a given amount of signal suppression occurs earlier than it would be with hard limiting. Another effect he shows is that, as limiting is lost, minor signal suppression occurs at relatively strong RSLs.

4. Experimental Verification

4.1 Having established the basic theoretical characteristics of FM receivers, a series of experiments were conducted to verify the theory and investigate effects not easily determined theoretically. In the experiments, two types of FM receiver were used. One receiver was a typical state-of-the-art LOS M/W receiver. This receiver was designed for normal operation with typical RSLs of -30 dBm and little fading. The characteristics of this receiver are listed on the next two pages. The IF response data was courtesy of J.R. Hammer, G.L. Cook, and R.J. Girvin. Unless otherwise mentioned, all data in this report captioned "LOS M/W Receiver" was taken using this receiver. The other receiver used was a typical TROPO M/W receiver. This receiver was designed to operate with a nominal RSL of -60 dBm. Due to the fading characteristics of TROPO RSL, the receiver was designed to operate satisfactorily with RSLs approaching FM threshold. The characteristics of this receiver are listed on the following pages. All data in this report captioned "TROPO M/W Receiver" was taken with this receiver.

4.2 The first experiment was to see if the previous baseband noise spectrum curves for hard and no limiting could be used to determine whether or not the limiter in a M/W receiver is limiting. Using a spectrum analyzer attached to the baseband of the M/W receiver, various levels of unmodulated RF carrier were applied to the receivers. The baseband noise spectrum as a function of input carrier to (thermal) noise (C/N) was photographed for both the LOS and the TROPO M/W receivers. The photographs (with image color reversed to facilitate reproduction) are on the next three pages. The noise spectrum was plotted in millivolts rather than milliwatts. However, since power is just voltage squared, the millivolt display is just compressed in vertical dimension relative to a milliwatt display. The shapes of either type of display is identical. Comparing the photographs with the "Hard Limiting" and "No Limiting" curves leave little doubt that the LOS M/W receiver has soft limiting for low C/Ns. However, both receivers have hard limiting for high C/Ns.

4.3 Next, the theoretical quieting curves were compared with actual quieting curves. This was done by measuring baseband noise with a frequency selective voltmeter while different unmodulated RF signals were applied to the M/W receiver. Baseband test tone suppression was also measured by applying a test tone (of frequency 0.05 times the IF bandwidth) to the M/W transmitter. The output of the transmitter was applied to the M/W receiver through an attenuator. Varying the amount of attenuation allowed various RSLs to be simulated.

4.4 The agreement between theory and experiment was excellent except for very low C/Ns. As mentioned in the theory section, with no carrier present at the receiver, the slot noise will be a direct function of IF limiting. Clearly neither the TROPO nor the LOS receiver is fully limiting at a -10 dB C/N. However, the TROPO receiver limiting is more

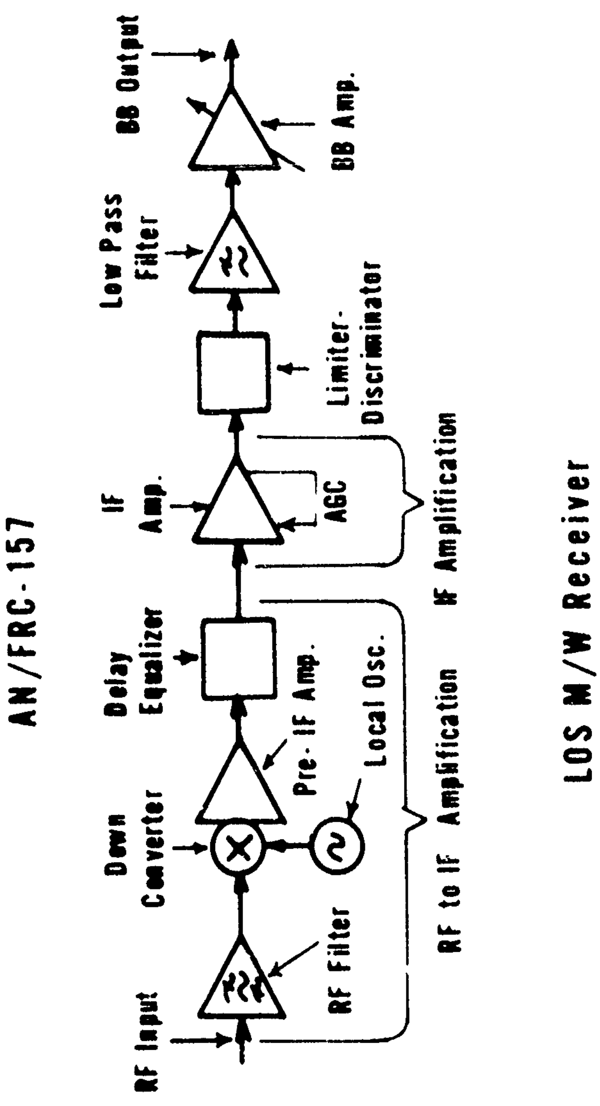

AN/FRC-157
RF Input
RF Filter
Down Converter
Local Osc.
Pre-IF Amp.
Delay Equalizer
IF Amp.
AGC
Limiter-Discriminator
Low Pass Filter
BB Amp.
BB Output
RF to IF Amplification
IF Amplification
LOS M/W Receiver

Figure 9

Channel Capacity: 600 VF Circuits

Baseband (BB) Range: 60 to 2660 kHz

Receive Carrier Frequency: 5 GHz

De-emphasis: None

Per Channel Deviation: 140 kHz (rms)

Noise Figure: 8.1 dB (overall)

RF to IF Amplification:

Overall Gain: +19 dB

Linear Amplification Range:
$-\infty$ to -17 dBm RF Input

1/2 dB Gain Compression:
-16 dBm RF Input

1 dB Gain Compression:
-15 dBm RF Input

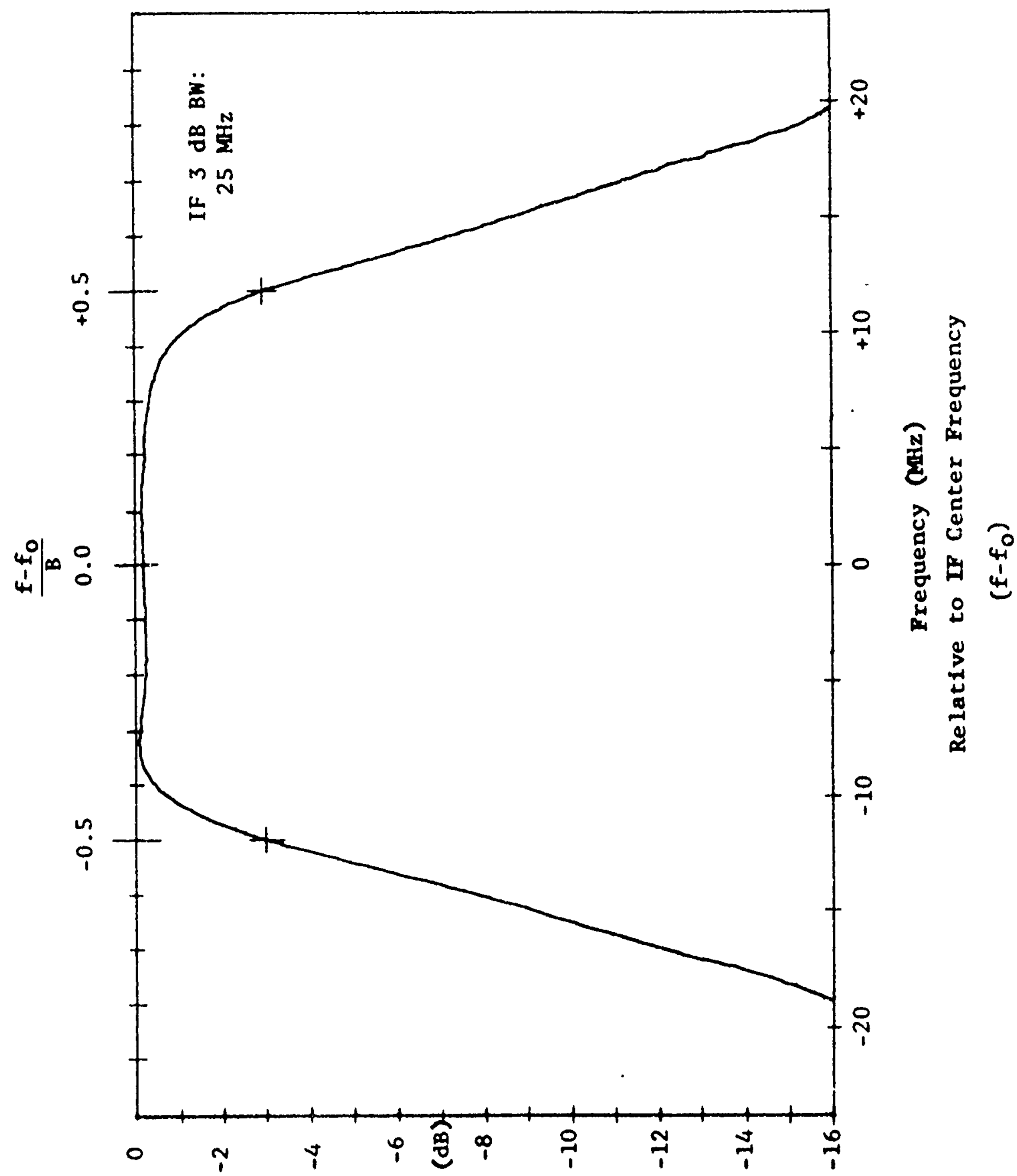
IF Amplification:

Maximum Gain: +40 dB

3 dB Bandwidth: 25 MHz

LOS M/W Receiver Characteristics

Table 12



LOS M/W Receiver IF Response

Figure 10

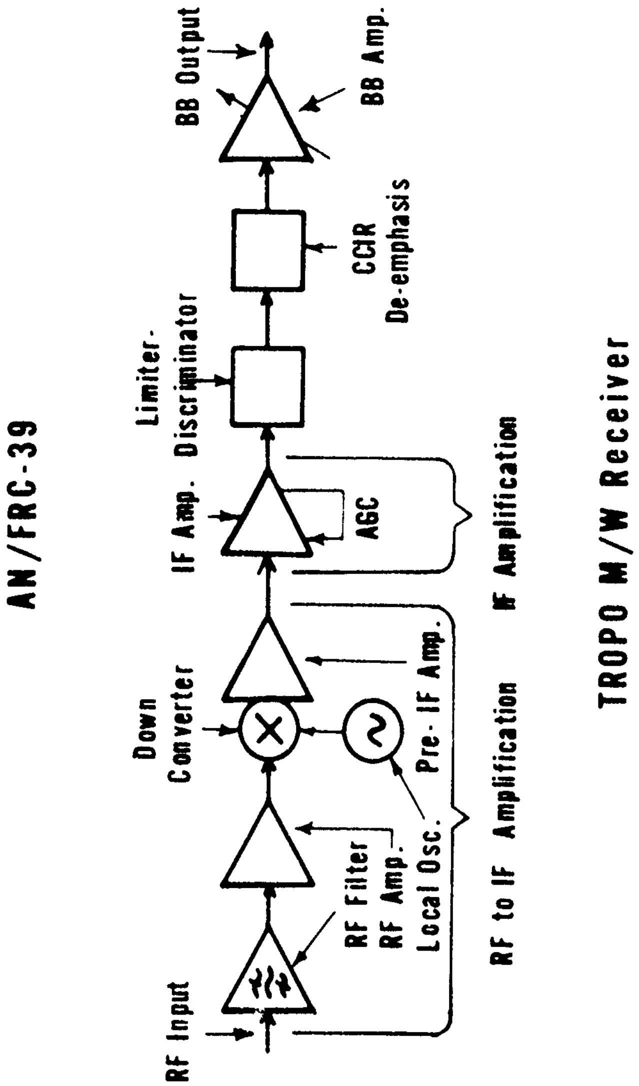

AN/FRC-39
RF Input
RF Filter
RF Amp.
Local Osc.
Down Converter
Pre- IF Amp.
IF Amp.
AGC
Limiter-Discriminator
CCIR De-emphasis
BB Amp.
BB Output
RF to IF Amplification
IF Amplification
TROPO M/W Receiver

Figure 11

Channel Capacity: 60 VF Circuits

Baseband (BB) Range: 12 to 252 kHz

Receive Carrier Frequency: 400 MHz

De-emphasis: CCIR

Per Channel Deviation: 100 kHz (rms)

Noise Figure: 1.5 dB (overall)

RF to IF Amplification:

Overall Gain: +40 dB

Linear Amplification Range:
$-\infty$ to -50 dBm RF Input

1/4 dB Gain Compression:
-40 dBm RF Input

1 dB Gain Compression:
-37 dBm RF Input

5 dB Gain Compression:
-30 dBm RF Input

15 dB Gain Compression:
-20 dB RF Input

IF Amplification:

Maximum Gain: +26 dB

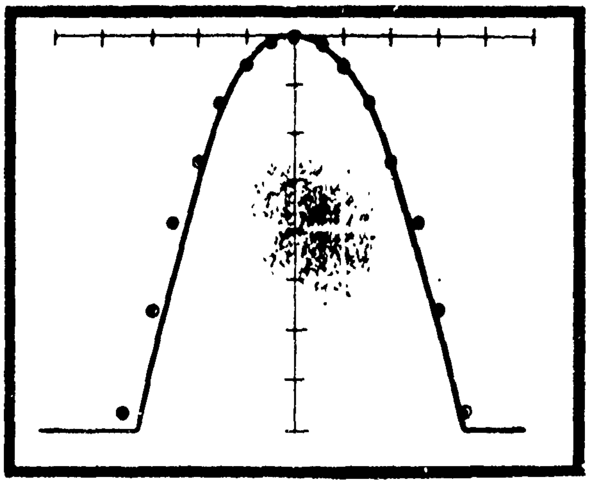
3 dB Bandwidth: 3.09 MHz

TROPO M/W Receiver Characteristics

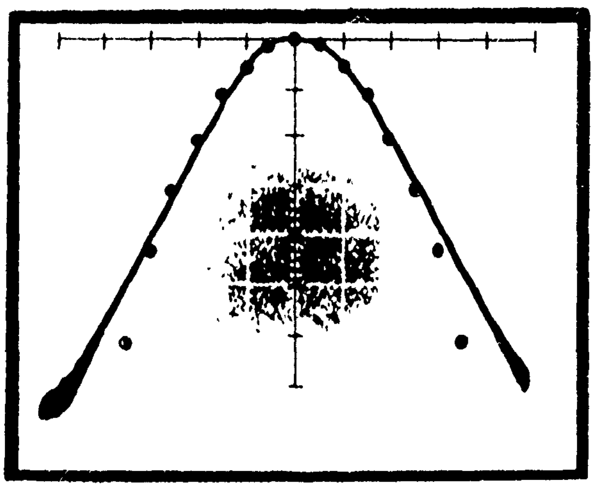
Table 13

IF Noise BW: 3.286 MHz
IF 3 dB BW: 3.087 MHz

curve - Actual IF Response
⊙ - Ideal Gaussian Response



Vertical: 2 dB/Div.
Horizontal: 1 MHz/Div.

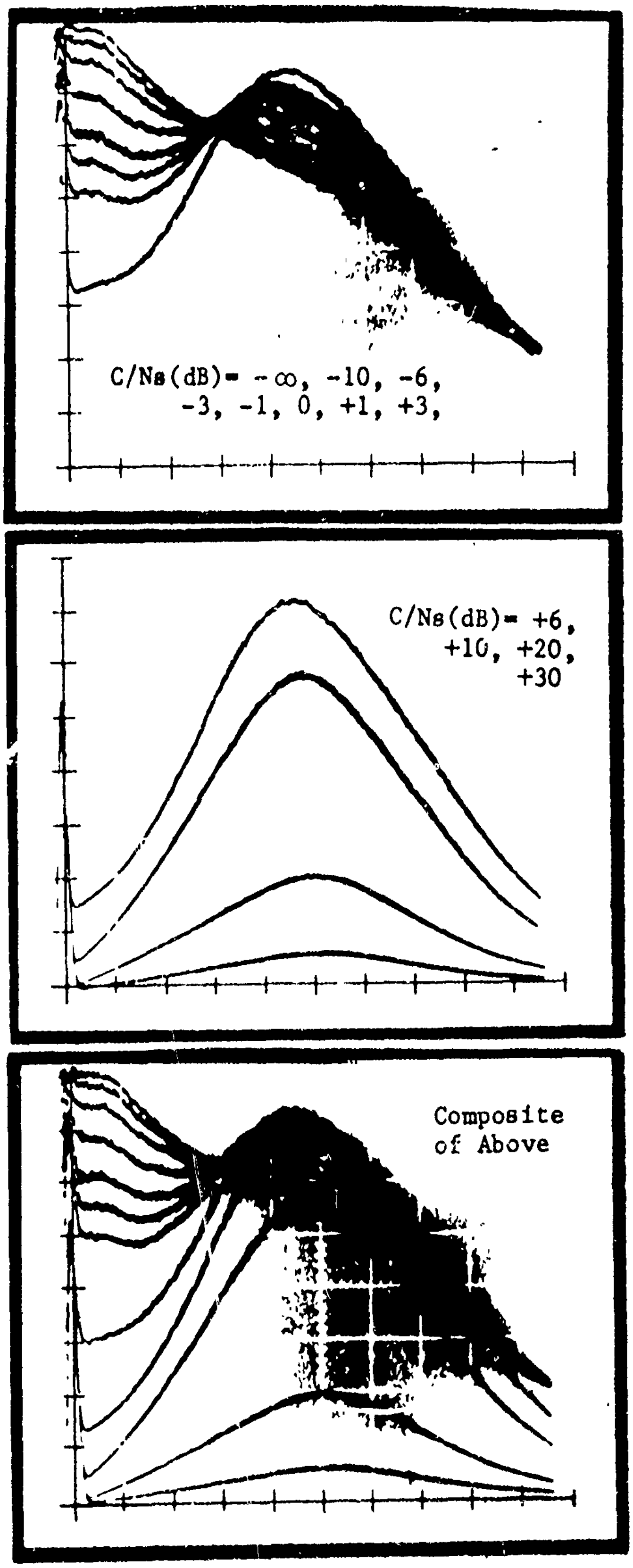


Vertical: 10 dB/Div.
Horizontal: 2 MHz/Div.

TROPO M/W Receiver IF Response

Figure 12

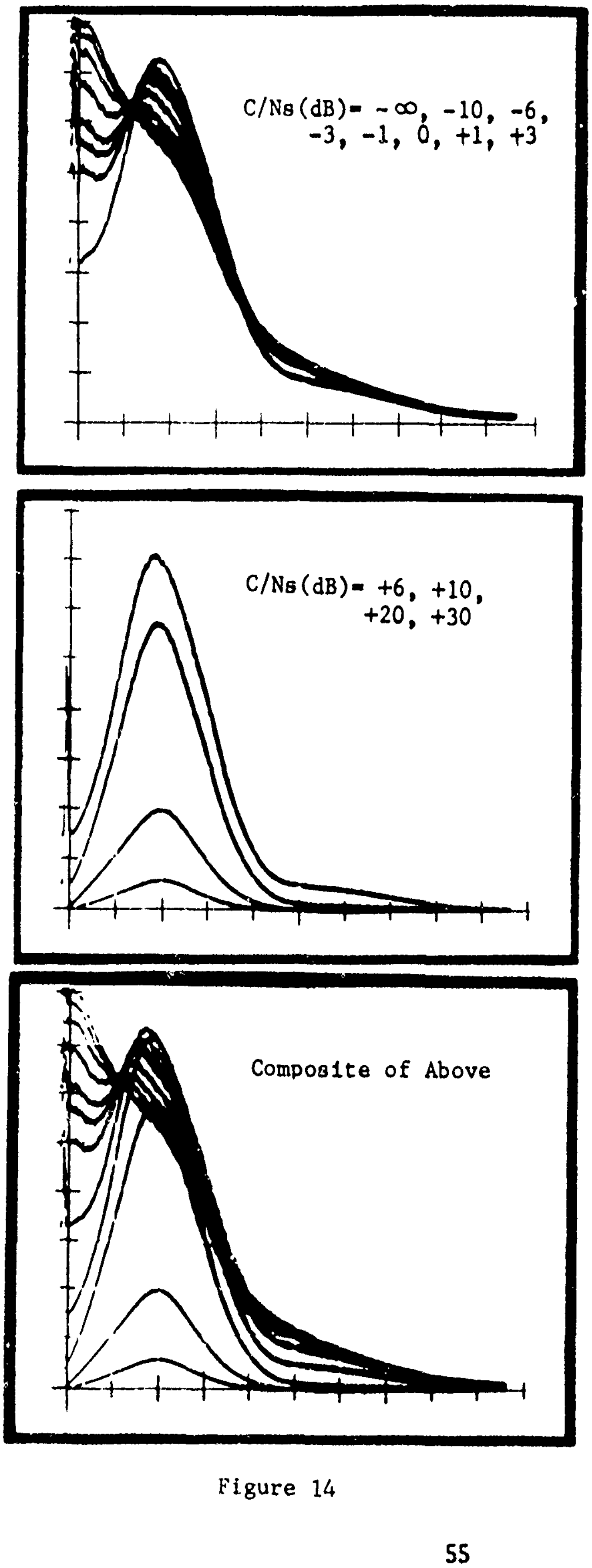
LOS M/W Receiver
no de-emphasis



Vertical: mV
Horizontal: 2 MHz/Div.
(0 to 20 MHz)

Baseband Noise Spectrum
for Various C/Ns

Figure 13

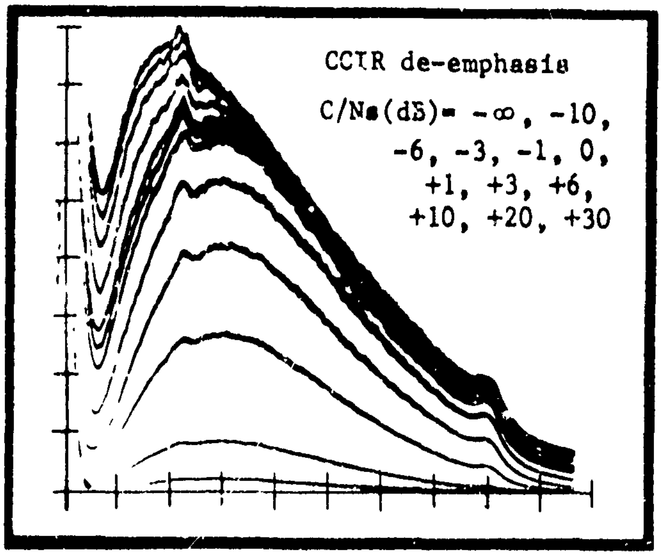


LOS M/W Receiver
no de-emphasis

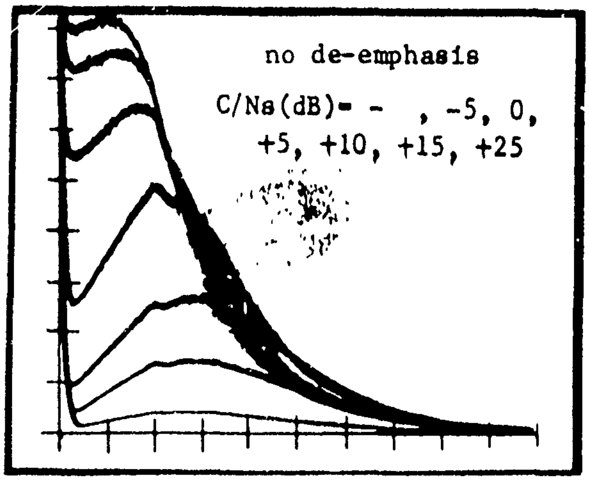
Vertical: mV
Horizontal: 5 MHz/Div.
(0 to 50 MHz)

Baseband Noise Spectrum
for Various C/Ns

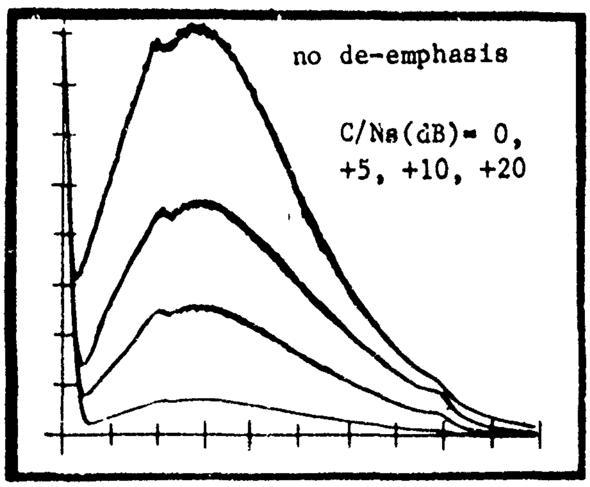
Figure 14



TROPO M/W Receiver



Vertical: mV
Horizontal: 500 kHz/Div.
(0 to 5 MHz)



Baseband Noise Spectrum
for Various C/Ns

Figure 15

effective (less soft) in the low C/N region. This is indicated by the TROPO slot noise values being closer to the theoretical curves than are the LOS noise measurements. This condition was also indicated by the previous experiment. Slot noise measurements were also taken with no signal into the M/W receivers. The results are plotted versus theoretical results on the next page.

4.5 In addition to the quieting curves of the normal receivers, the experiment was repeated with varying degrees of limiter action. To reduce the limiting action in the receivers, various degrees of attenuation (attenuator pads) were placed between the receiver mixer (down converter) and the input to the IF amplifier. The IF amplifier is a variable gain amplifier which tries to keep the input level to the limiter section constant. Placing the attenuation (pad) in front of the IF amplifier causes the amplifier to increase its gain to compensate for the pad loss. As attenuation is increased, due to limited gain, the IF amplifier will be less effective in holding the signal constant into the limiter section. As the level into the limiter section is reduced, limiting action is reduced (goes soft). When the pad is placed in the receiver, not only is limiter action affected but noise figure is also increased. Compensation was made for changes in noise figure in the graphed data. Also, the TROPO receiver data has been corrected to delete the frequency response of the CCIR de-emphasis network.

4.6 Notice in the following curves that as limiter action goes soft, the noise for very low C/Ns is reduced. However, the noise in the low frequency baseband slots start to experience additional noise near FM threshold (approximately +10 dB C/N). For moderate amounts of limiter degradation, about the only effect on noise in the FM threshold region is to cause the low slot noise to increase more rapidly than normal. In extreme cases, however, even high frequency noise slots are affected. Also, note that the LOS receiver is significantly more sensitive to pre-IF signal attenuation than is the TROPO receiver. The previous data has also been plotted relative to the normal configuration data. This shows the effect of both limiter and overall noise figure degradation.

4.7 Since the data indicated that the limiter of the LOS M/W receiver was soft at low C/N ratios, a wideband amplifier was placed at the input to the IF amplifier (the same place the pads had been placed) to drive the limiter harder for the low C/Ns. The quieting curve was reaccomplished and, as expected, FM threshold moved to the left and slot noise with no carrier present was increased. The following two pages compare theoretical and measured 1 dB FM noise thresholds and 20 dB noise quieting valves.

4.8 Several **sources have** indicated that if a received FM signal is modulated, slot noise will increase (relative to slot noise produced by an unmodulated carrier) near FM threshold. Middleton indicated that the slot noise would be even greater without limiting. These effects were investigated by white noise loading the LOS M/W transmitter with a Noise Power Ratio transmitter connected to the baseband input to the transmitter. The

FM Receiver Baseband Slot Noise
(no signal input, no de-emphasis)

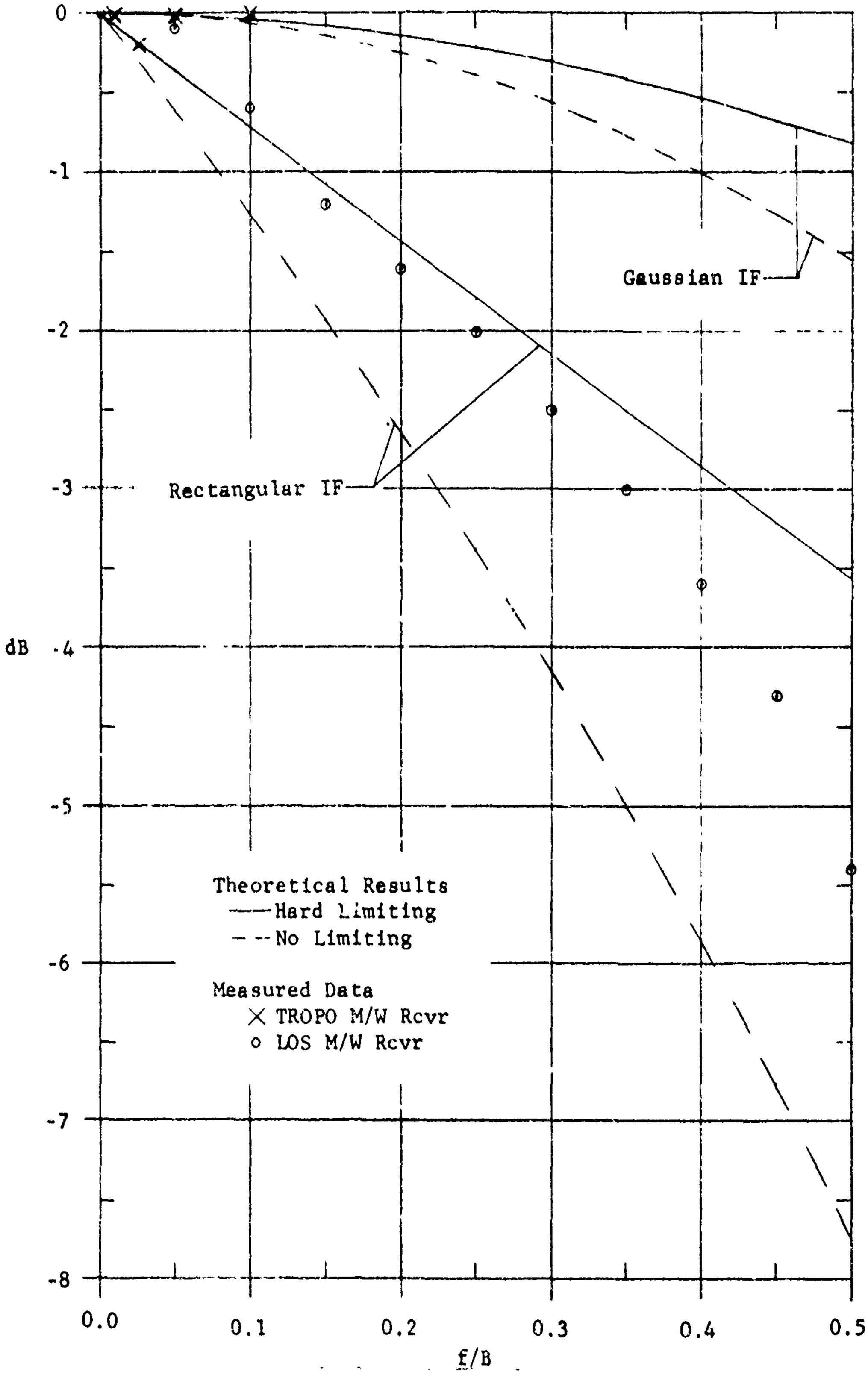
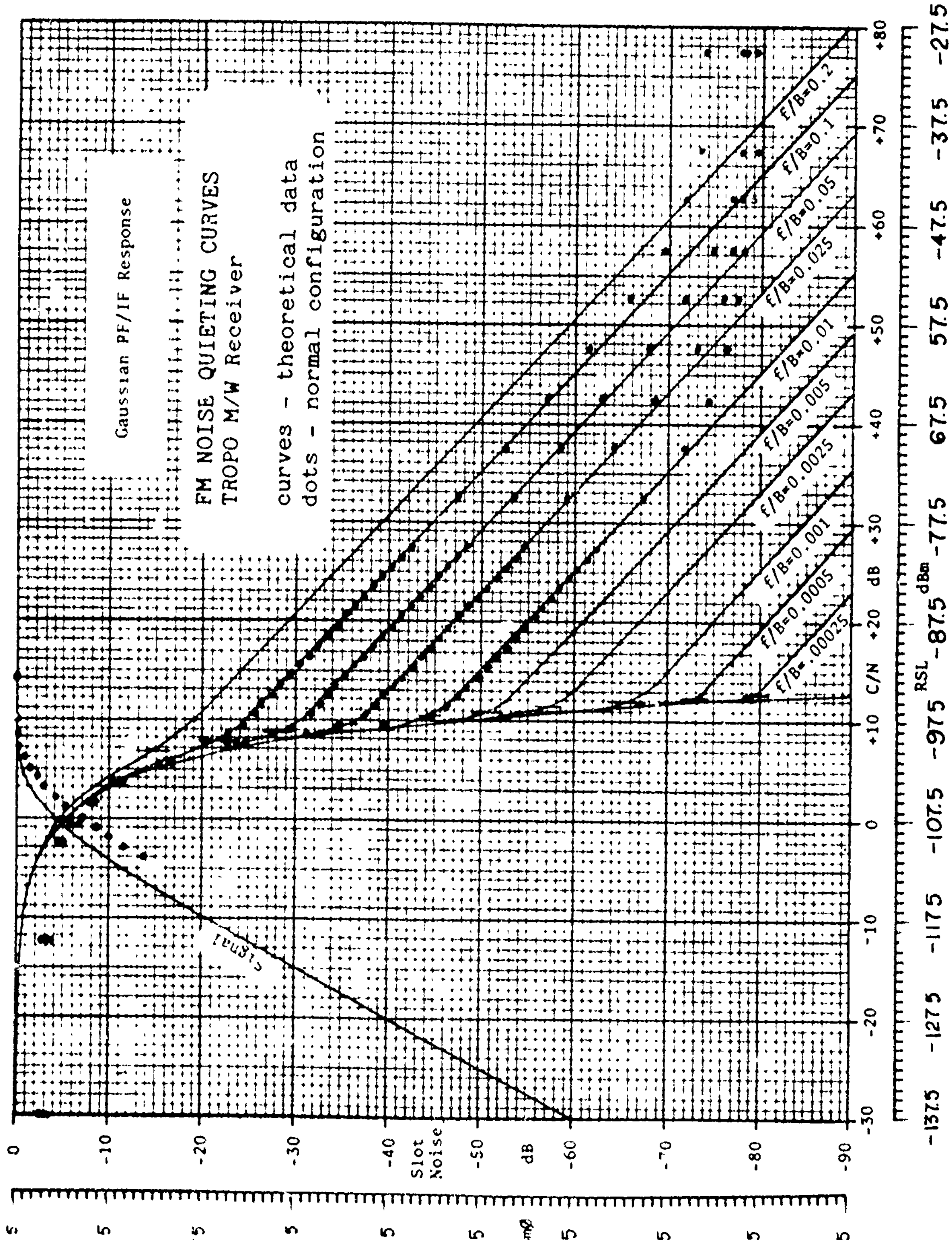


Figure 15.1



Gaussian PF/IF Response
FM NOISE QUIETING CURVES
TROPO M/W Receiver
curves - theoretical data
dots - normal configuration
f/B=0.2
f/B=0.1
f/B=0.05
f/B=0.025
f/B=0.01
f/B=0.005
f/B=0.0025
f/B=0.001
f/B=0.0005
f/B=.00025
Signal
C/N dB
RSL dBm
Slot Noise dB
dBm0

Figure 16

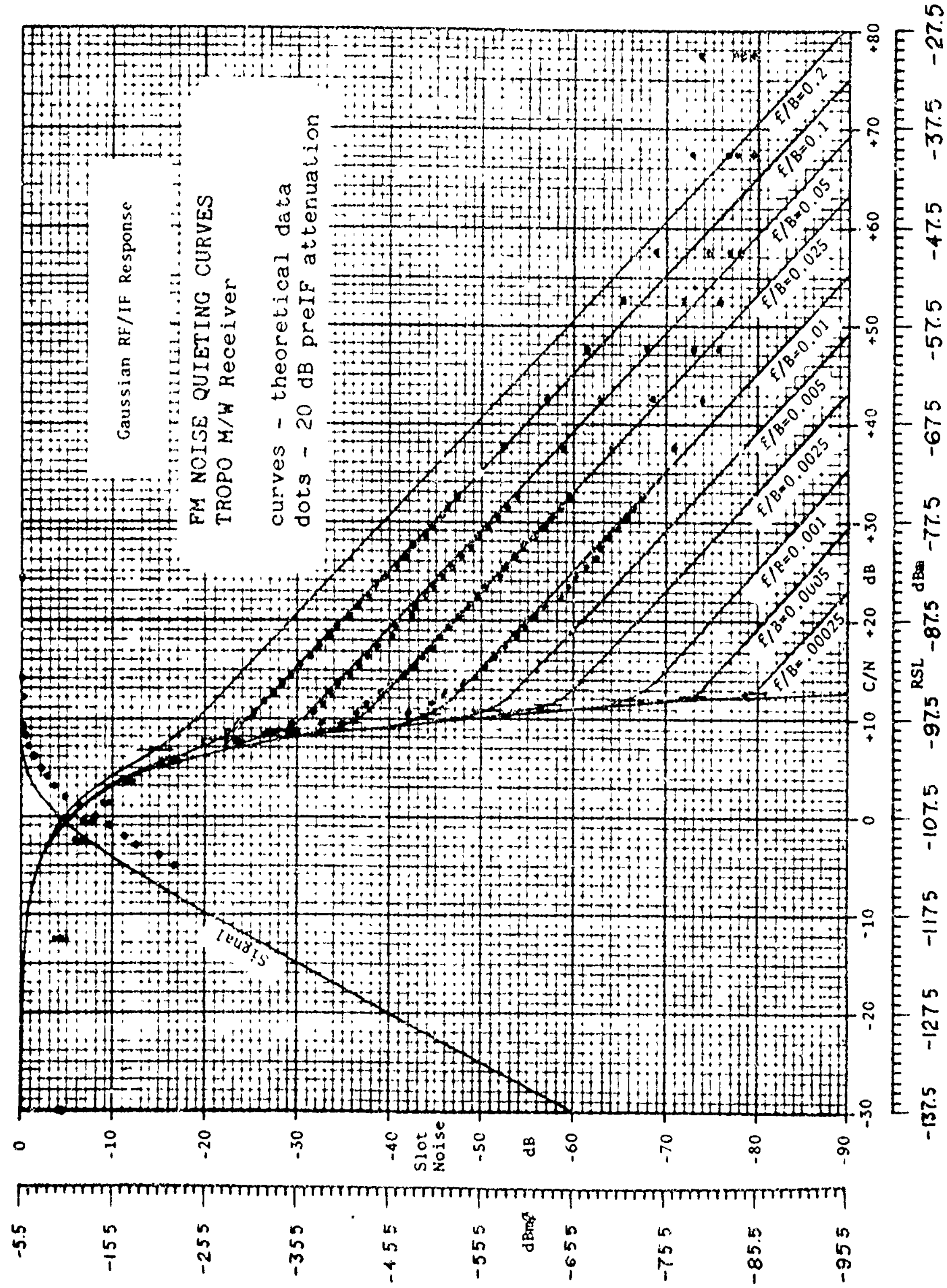
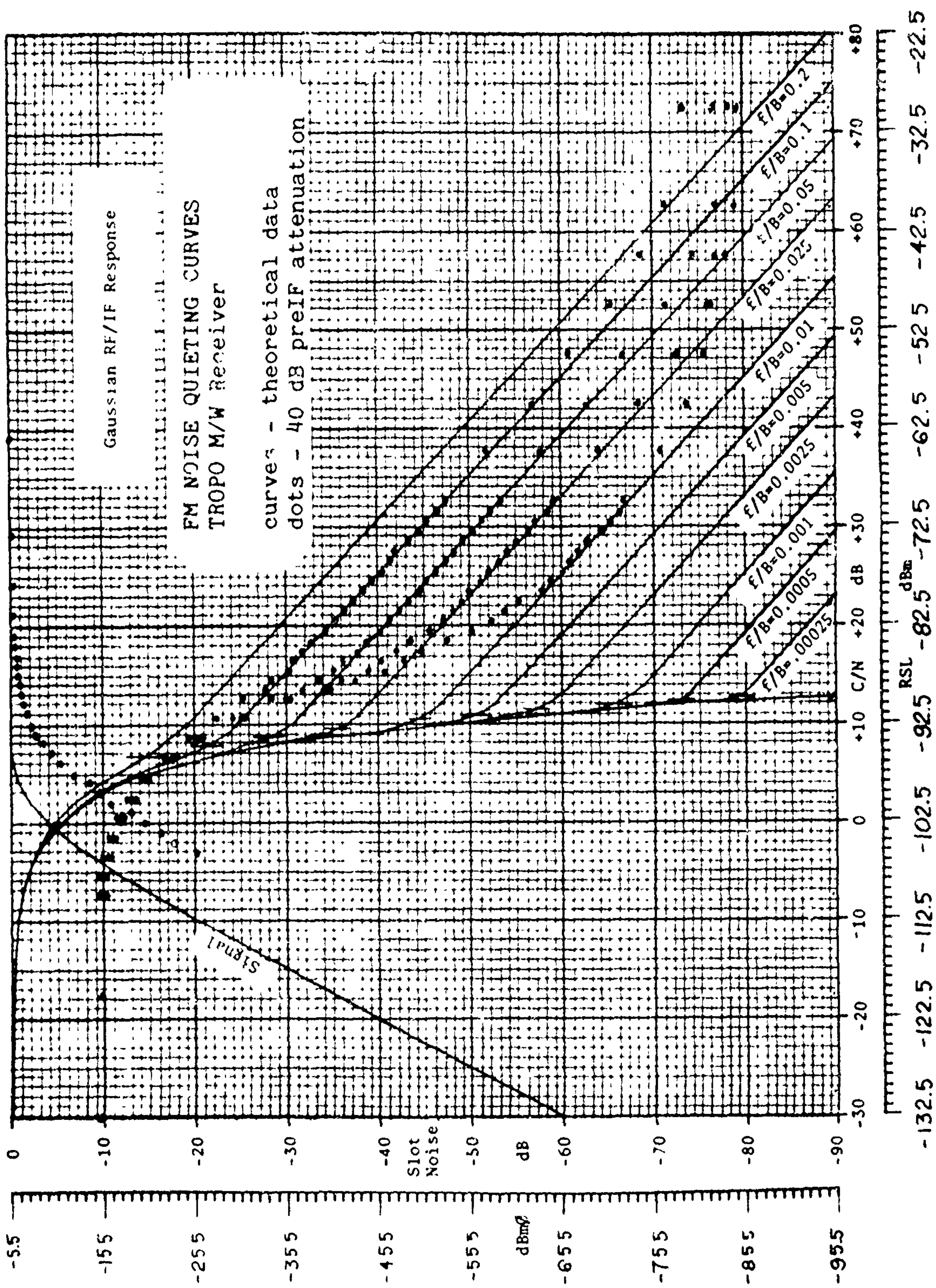

Gaussian RF/IF Response
FM NOISE QUIETING CURVES
TROPO M/W Receiver
curves - theoretical data
dots - 20 dB preIF attenuation
Signal
f/B=0.2
f/B=0.1
f/B=0.05
f/B=0.025
f/B=0.01
f/B=0.005
f/B=0.0025
f/B=0.001
f/B=0.0005
f/B=.00025
C/N dB
RSL dBm
Slot Noise dB
dBm0

Figure 17



FM NOISE QUIETING CURVES
TROPO M/W Receiver
curves - theoretical data
dots - 40 dB preIF attenuation
Gaussian RF/IF Response
Signal
f/B=0.2
f/B=0.1
f/B=0.05
f/B=0.025
f/B=0.01
f/B=0.005
f/B=0.0025
f/B=0.001
f/B=0.0005
f/B=.00025
C/N dB
RSL dBm
Slot Noise dB
dBm0

Figure 18

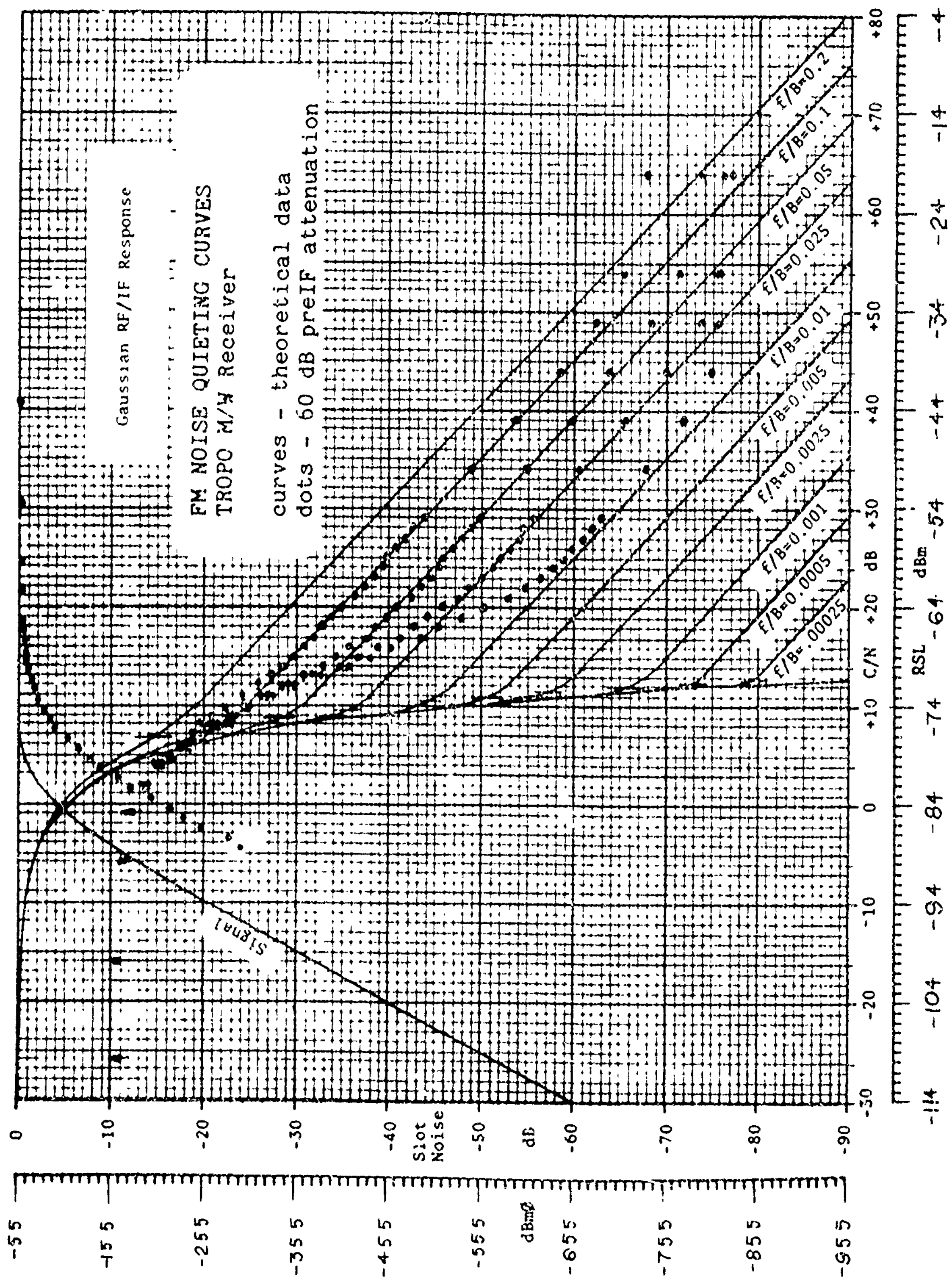

Gaussian RF/IF Response
FM NOISE QUIETING CURVES
TROPO M/W Receiver
curves - theoretical data
dots - 60 dB preIF attenuation
Signal
f/B=0.2
f/B=0.1
f/B=0.05
f/B=0.025
f/B=0.01
f/B=0.005
f/B=0.0025
f/B=0.001
f/B=0.0005
f/B=.00025
C/N dB
RSL dBm
Slot Noise dB
dBm0

Figure 19

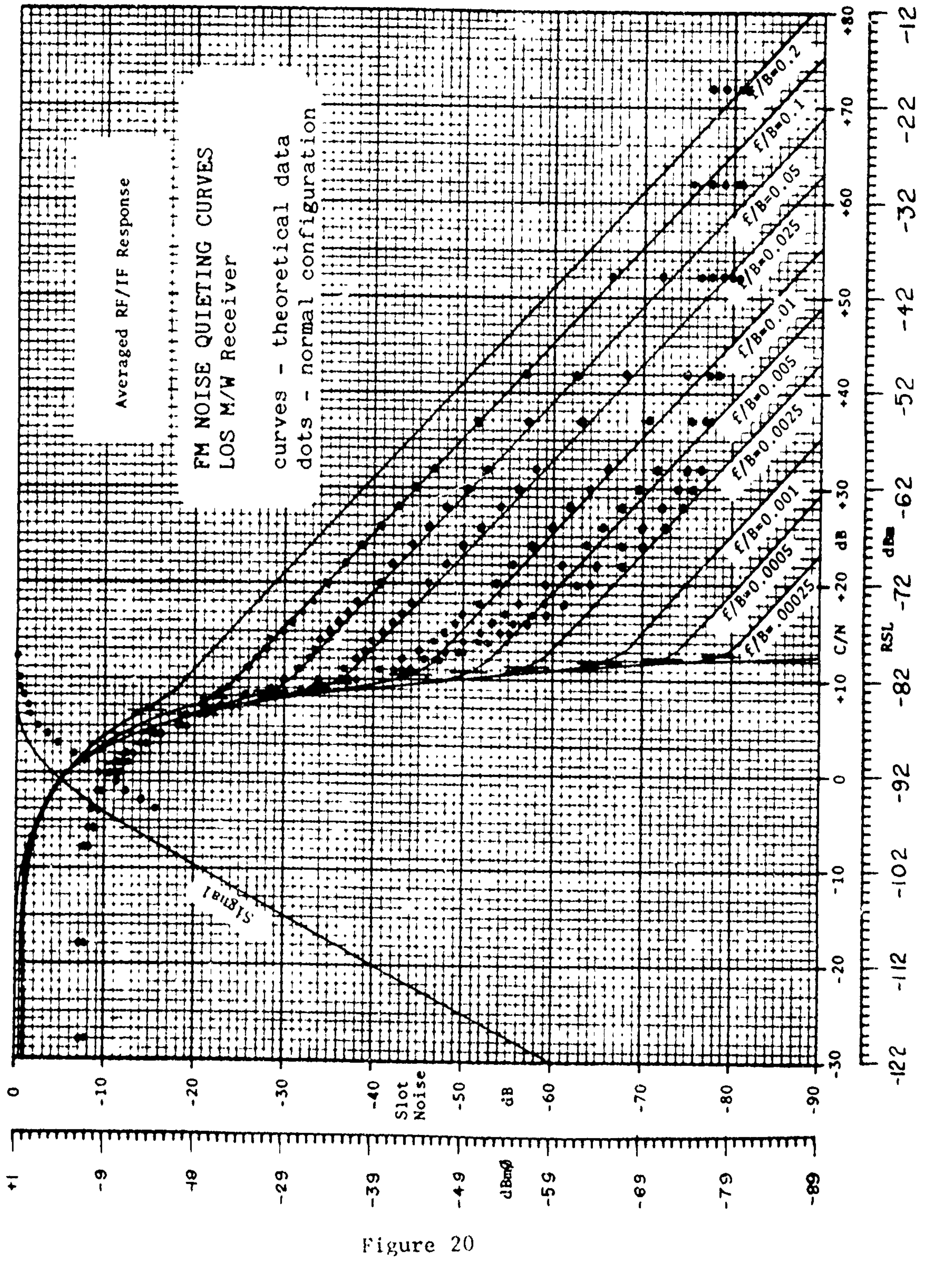

Averaged RF/IF Response
FM NOISE QUIETING CURVES
LOS M/W Receiver
curves - theoretical data
dots - normal configuration
Signal
f/B=0.2
f/B=0.1
f/B=0.05
f/B=0.025
f/B=0.01
f/B=0.005
f/B=0.0025
f/B=0.001
f/B=0.0005
f/B=.00025
Slot Noise dB
dBm0
C/N dB
RSL dBm

Figure 20

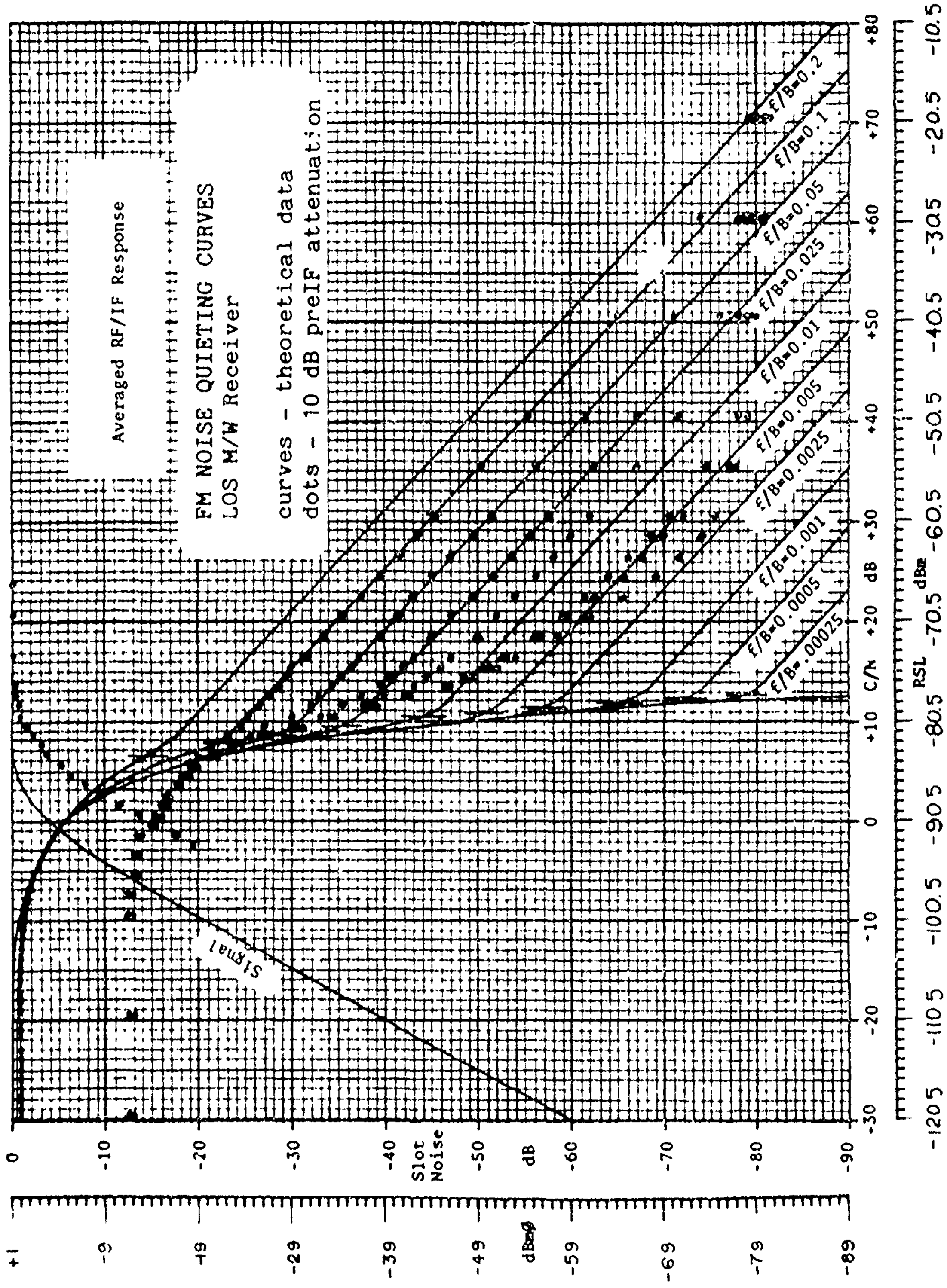

FM NOISE QUIETING CURVES
LOS M/W Receiver
curves - theoretical data
dots - 10 dB preIF attenuation
Averaged RF/IF Response
Signal
f/B=0.2
f/B=0.1
f/B=0.05
f/B=0.025
f/B=0.01
f/B=0.005
f/B=0.0025
f/B=0.001
f/B=0.0005
f/B=.00025
Slot Noise dB
C/N dB
RSL dBm
dBm0

Figure 21

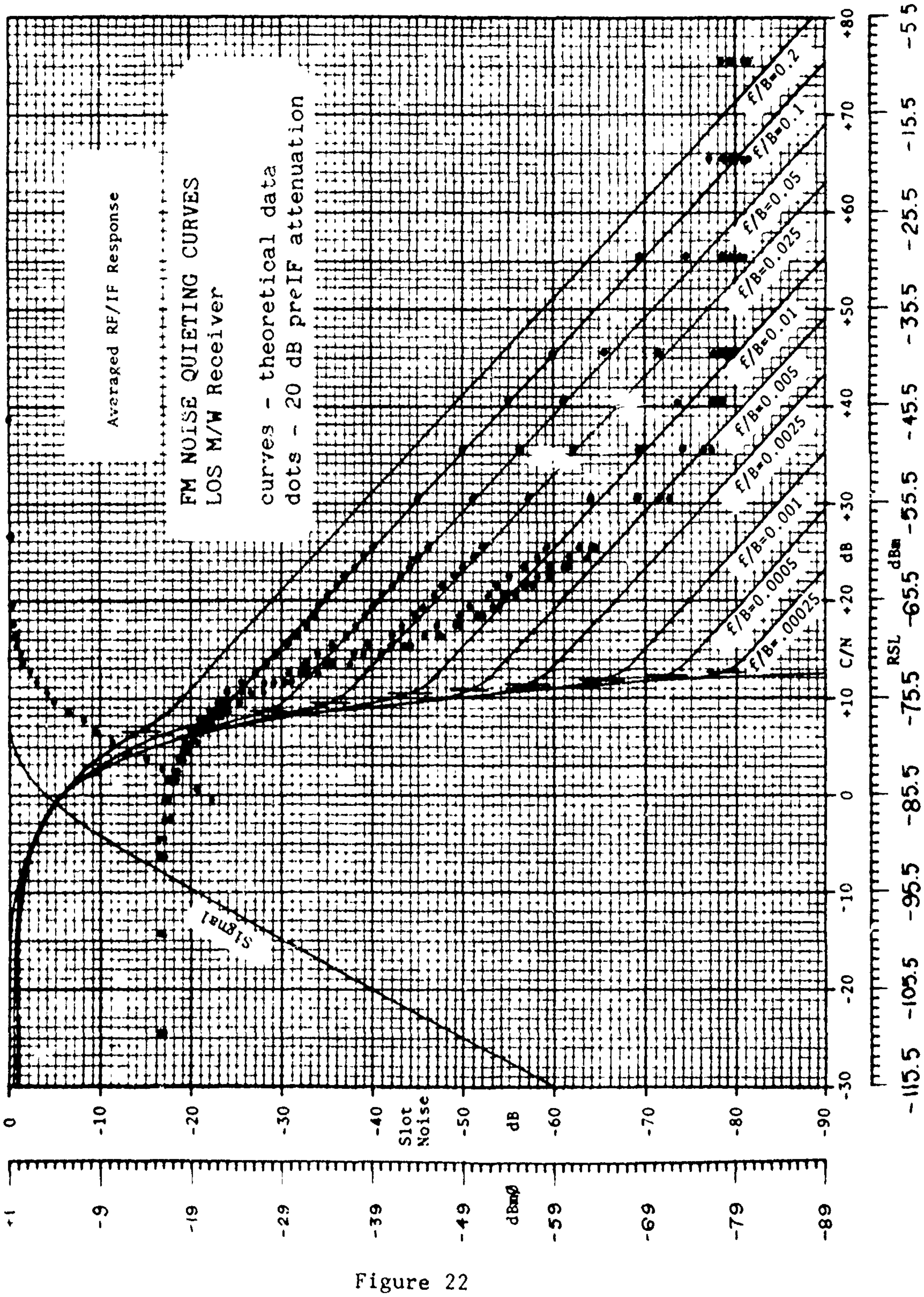
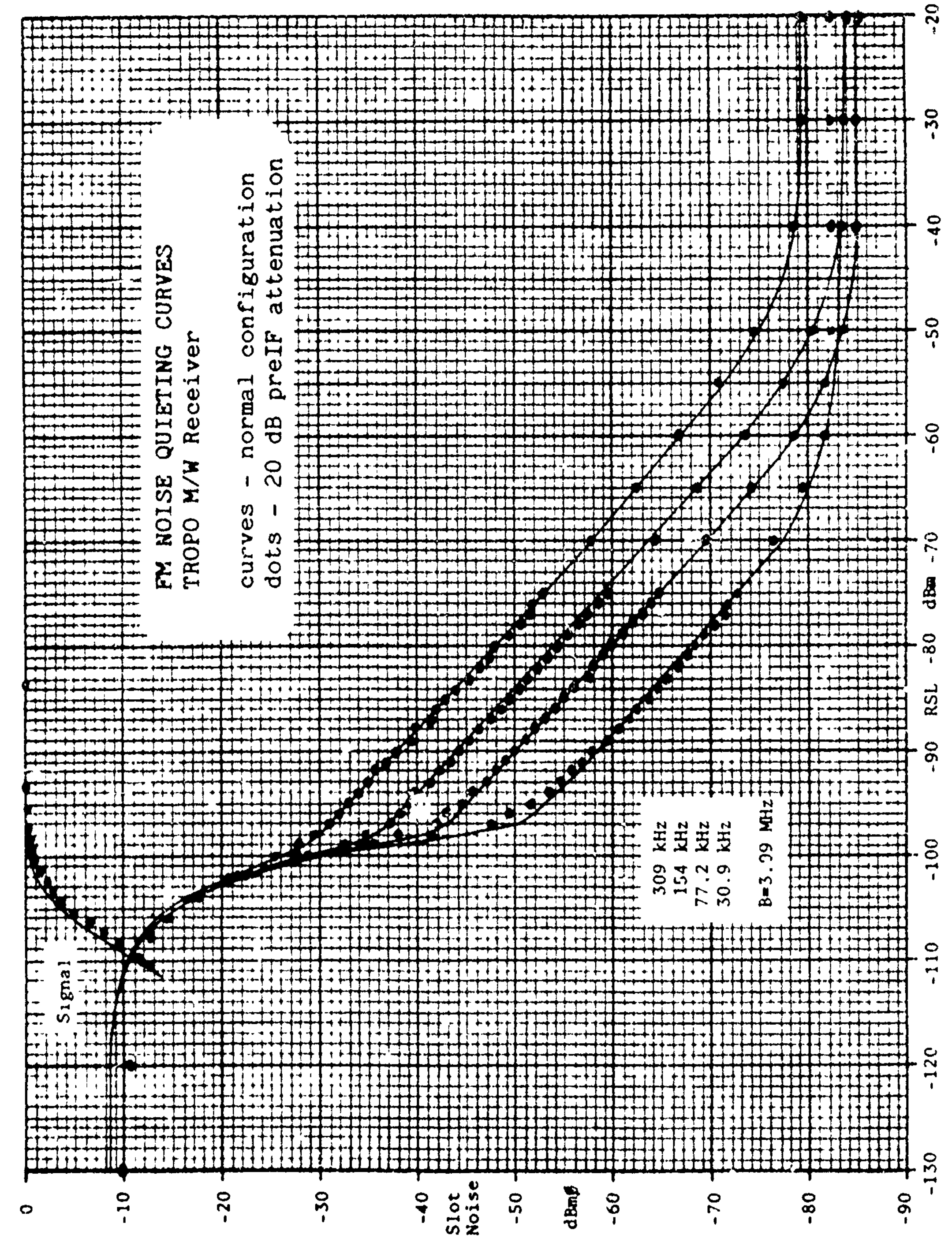

FM NOISE QUIETING CURVES
LOS M/W Receiver
curves - theoretical data
dots - 20 dB preIF attenuation
Averaged RF/IF Response
Signal
f/B=0.2
f/B=0.1
f/B=0.05
f/B=0.025
f/B=0.01
f/B=0.005
f/B=0.0025
f/B=0.001
f/B=0.0005
f/B=.00025
C/N dB
RSL dBm
Slot Noise dB
dBmØ

Figure 22



FM NOISE QUIETING CURVES
TROPO M/W Receiver
curves - normal configuration
dots - 20 dB preIF attenuation
309 kHz
154 kHz
77.2 kHz
30.9 kHz
B=3.99 MHz
Signal
Slot Noise
dBm0
RSL dBm
-130 -120 -110 -100 -90 -80 -70 -60 -50 -40 -30 -20
0 -10 -20 -30 -40 -50 -60 -70 -80 -90

Figure 23

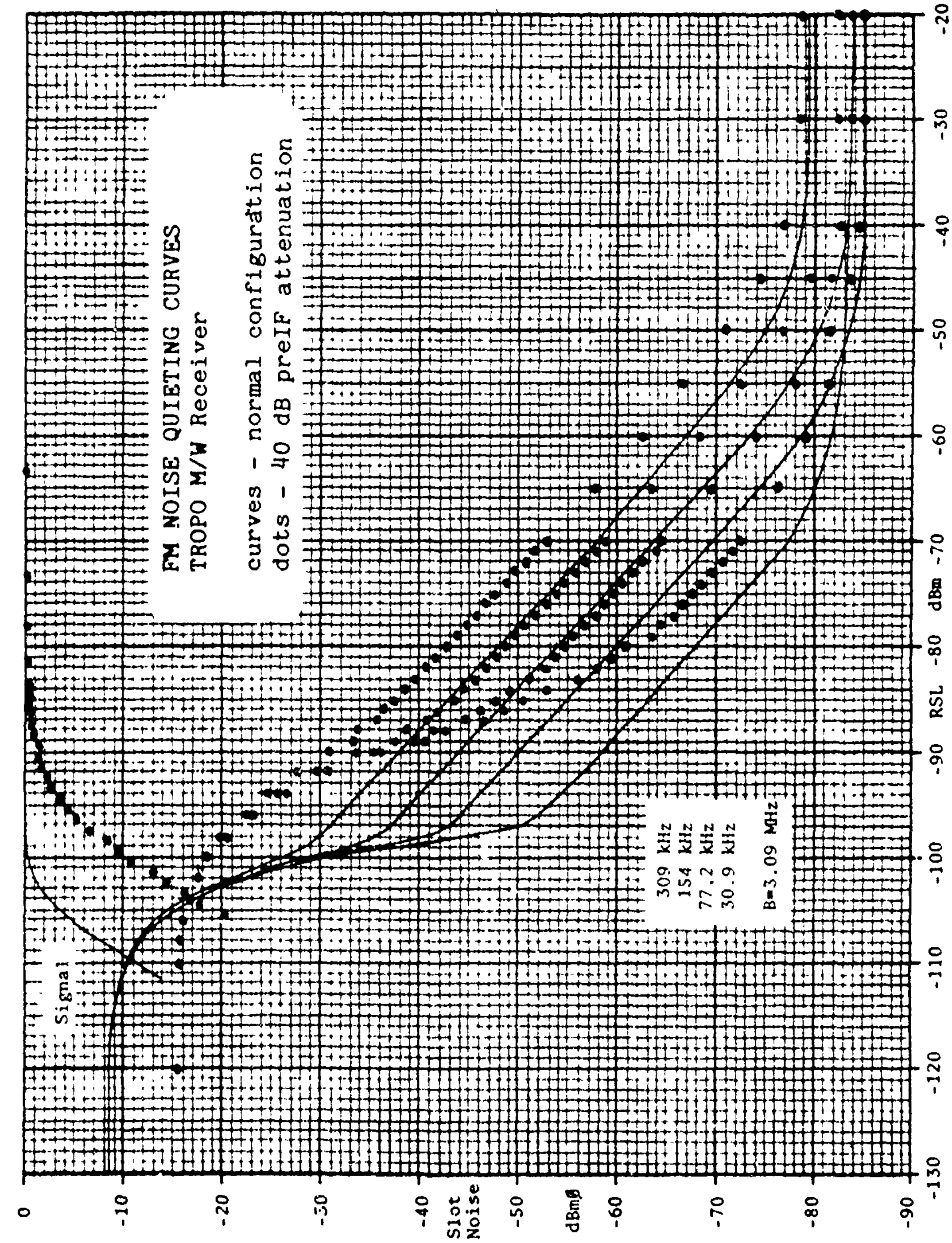

FM NOISE QUIETING CURVES
TROPO M/W Receiver
curves - normal configuration
dots - 40 dB preIF attenuation
Signal
309 kHz
154 kHz
77.2 kHz
30.9 kHz
B=3.09 MHz
RSL dBm
Slot Noise
dBm0

Figure 24

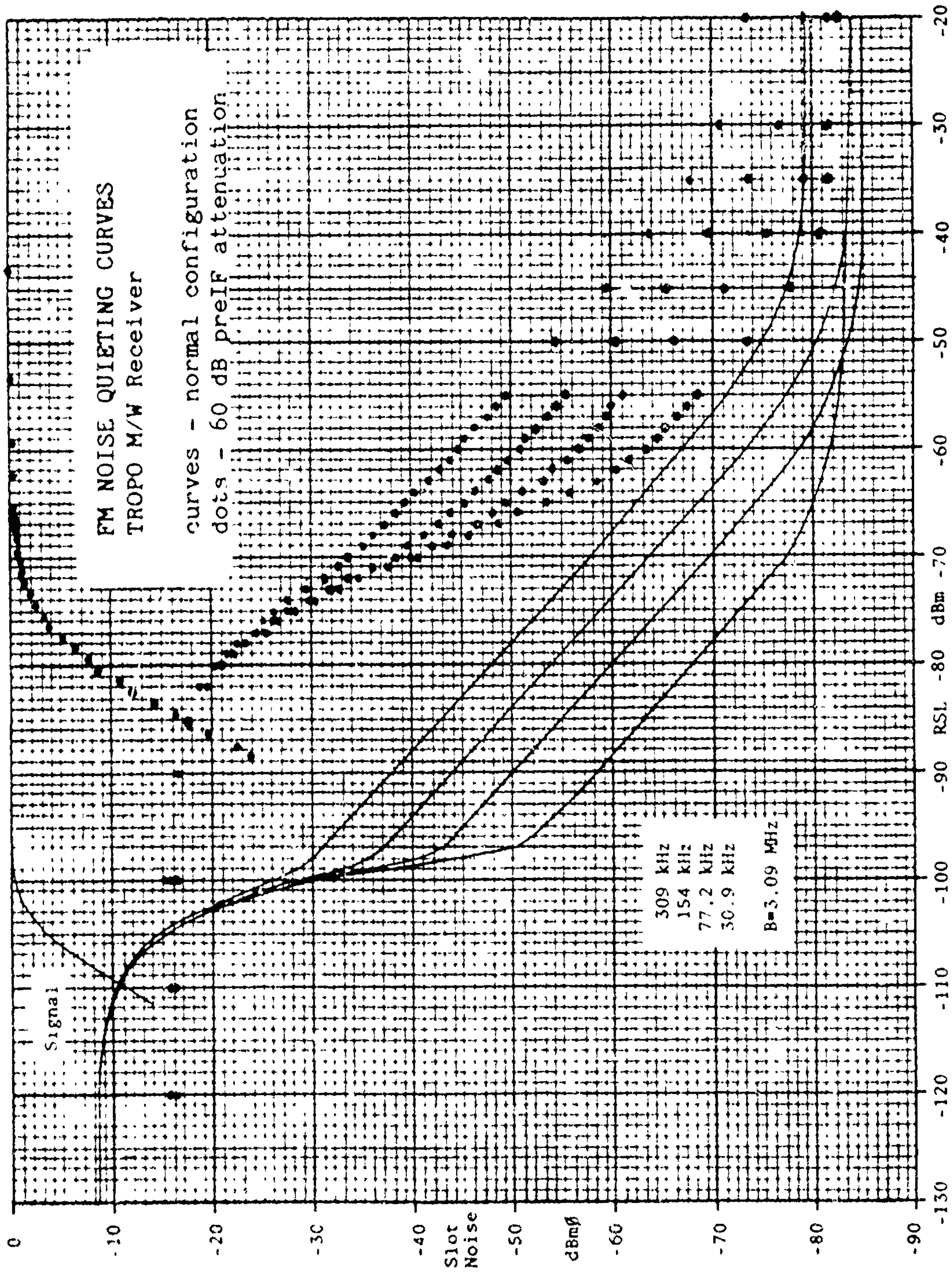

FM NOISE QUIETING CURVES
TROPO M/W Receiver
curves - normal configuration
dots - 60 dB preIF attenuation
Signal
309 kHz
154 kHz
77.2 kHz
30.9 kHz
B=3.09 MHz
RSL dBm
-130 -120 -110 -100 -90 -80 -70 -60 -50 -40 -30 -20
Slot Noise dBm0
0 -10 -20 -30 -40 -50 -60 -70 -80 -90

Figure 25

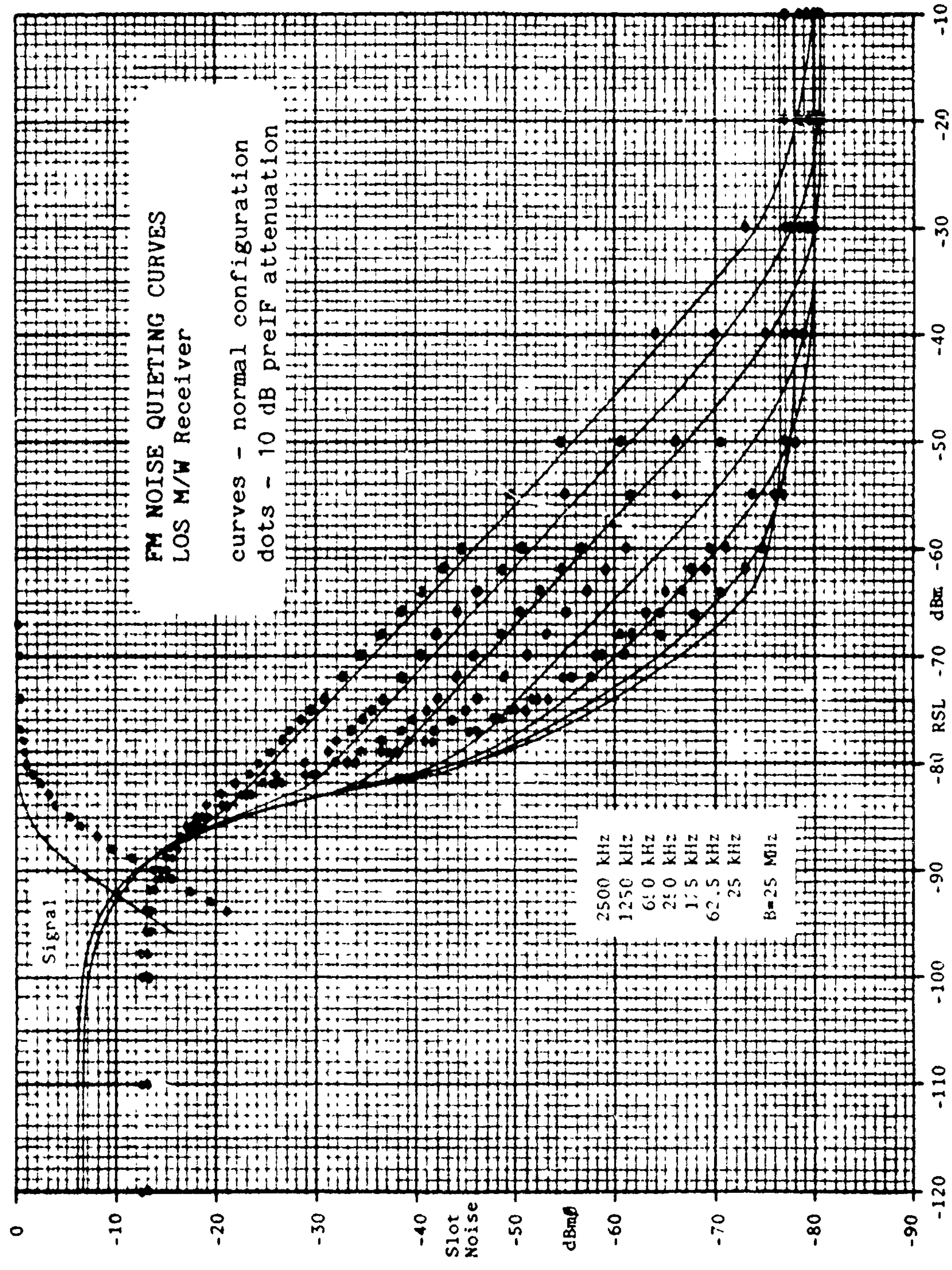

FM NOISE QUIETING CURVES
LOS M/W Receiver
curves - normal configuration
dots - 10 dB preIF attenuation
Signal
2500 kHz
1250 kHz
25 kHz
B=25 MHz
Slot Noise
dBm0
RSL
dBm
0
-10
-20
-30
-40
-50
-60
-70
-80
-90
-120
-110
-100
-90
-80
-70
-60
-50
-40
-30
-20
-10

Figure 26

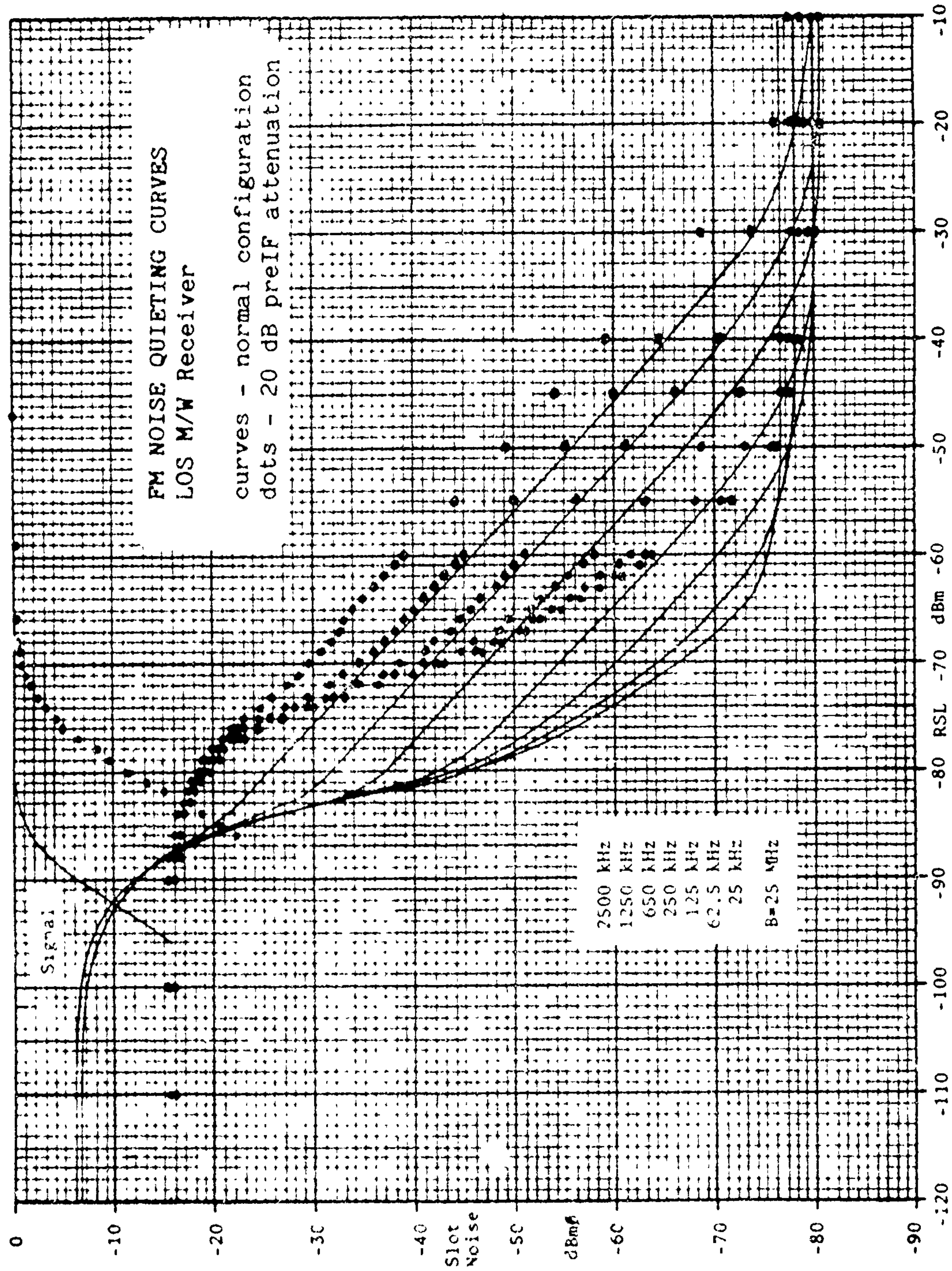
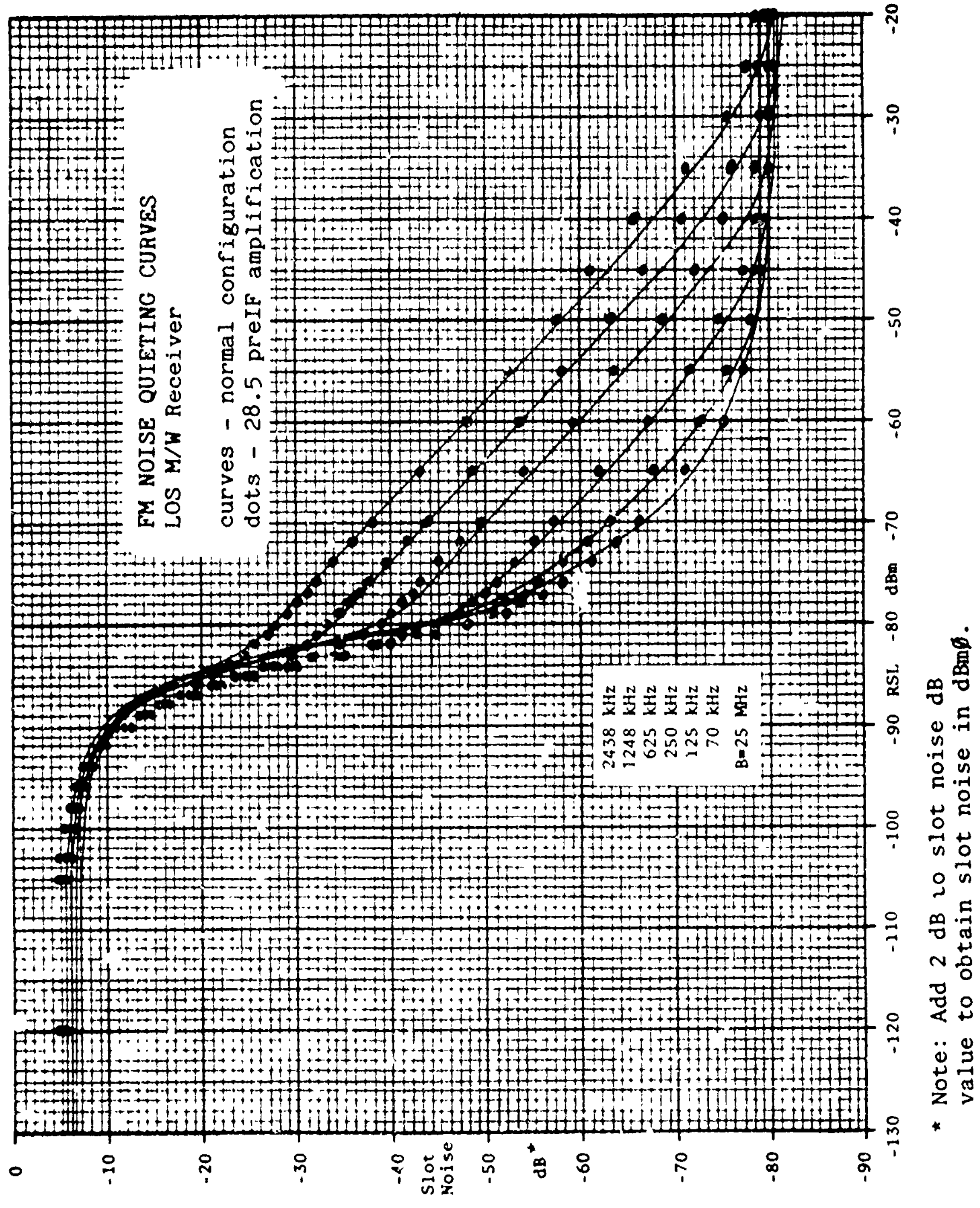

FM NOISE QUIETING CURVES
LOS M/W Receiver
curves - normal configuration
dots - 20 dB preIF atenuation
2500 kHz
1250 kHz
650 kHz
250 kHz
125 kHz
62.5 kHz
25 kHz
B=25 MHz
Signal
RSL dBm
Slot Noise dBm
0
-10
-20
-30
-40
-50
-60
-70
-80
-90
-100
-110
-120

Figure 27



FM NOISE QUIETING CURVES
LOS M/W Receiver
curves - normal configuration
dots - 28.5 preIF amplification
2438 kHz
1248 kHz
625 kHz
250 kHz
125 kHz
70 kHz
B=25 MHz
Slot Noise dB*
RSL dBm
0
-10
-20
-30
-40
-50
-60
-70
-80
-90
-130
-120
-110
-100
* Note: Add 2 dB to slot noise dB value to obtain slot noise in dBm0.

Figure 28

TROPO M/W Receiver

C/N (dB)

f/B	Gaus. IF	normal	20 dB pad	40 dB pad	60 dB pad
0.1	7.9	7.5	7.5	14.5	12
0.05	8.8	8.5	9.5	15.5	16
0.025	9.6	9.5	9.5	18.5	20
0.01	10.4	9.5	12.5	22.5	22
	Theoretical	Measured			

LOS M/W Receiver

C/N (dB)

f/B	Avg. IF	30 dB gain	normal	10 dB pad	20 dB pad
0.1	7.7	8	7	7.5	11.5
0.05	8.7	9	8	10.5	14.5
0.025	9.5	11	10	12.5	15.5
0.01	10.3	13	13	13.5	22.5
	Theoretical	Measured			

FM Noise Threshold

Table 14

TROPO M/W Receiver

C/N (dB)

f/B	Gaus. IF	normal	20 dB pad	40 dB pad	60 dB pad
0.1	7.2	8.5	9.5	15.5	16
0.05	6.5	9.5	11	18	19
0.025	6.4	9.5	11.5	18.5	19.5
0.01	6.3	9.5	11.5	18.5	19.5
	Theoretical	Measured			

LOS M/W Receiver

C/N (dB)

f/B	Avg. IF	30 dB gain	normal	10 dB pad	20 dB pad
0.1	7.2	10	12.5	18.5	22.5
0.05	6.4	12.5	16.5	24.5	28.5
0.025	6.3	12.5	17	26.5	31
0.01	6.2	13	17	26.5	31.5
	Theoretical	Measured			

20 dB Noise Quieting

Table 15

modulated RF signal was applied to the LOS receiver using a variable attenuator. Basic Intrinsic Noise Ratio (BINR) and Noise Power Ratio (NPR) measurements were taken using a Noise Power Ratio receiver connected to the baseband of the M/W receiver (See Tant's book for an explanation of this measurement.). The BINR and NPR measurements were converted to slot noise measurements. Theoretical results were determined by using Rice's results for white Gaussian noise loading and the Anuff/Lou formula for required IF bandwidth. The results were calculated for CCIR loading and a baseband to IF bandwidth ratio of 0.1. The additional assumption was made that the additional noise due to Gaussian baseband noise loading was essentially flat spectrum over the radio baseband and equal to the noise spectral density at zero frequency. The results are tabulated and graphed on the next pages. The increased slot noise due to lack of limiting can be observed.

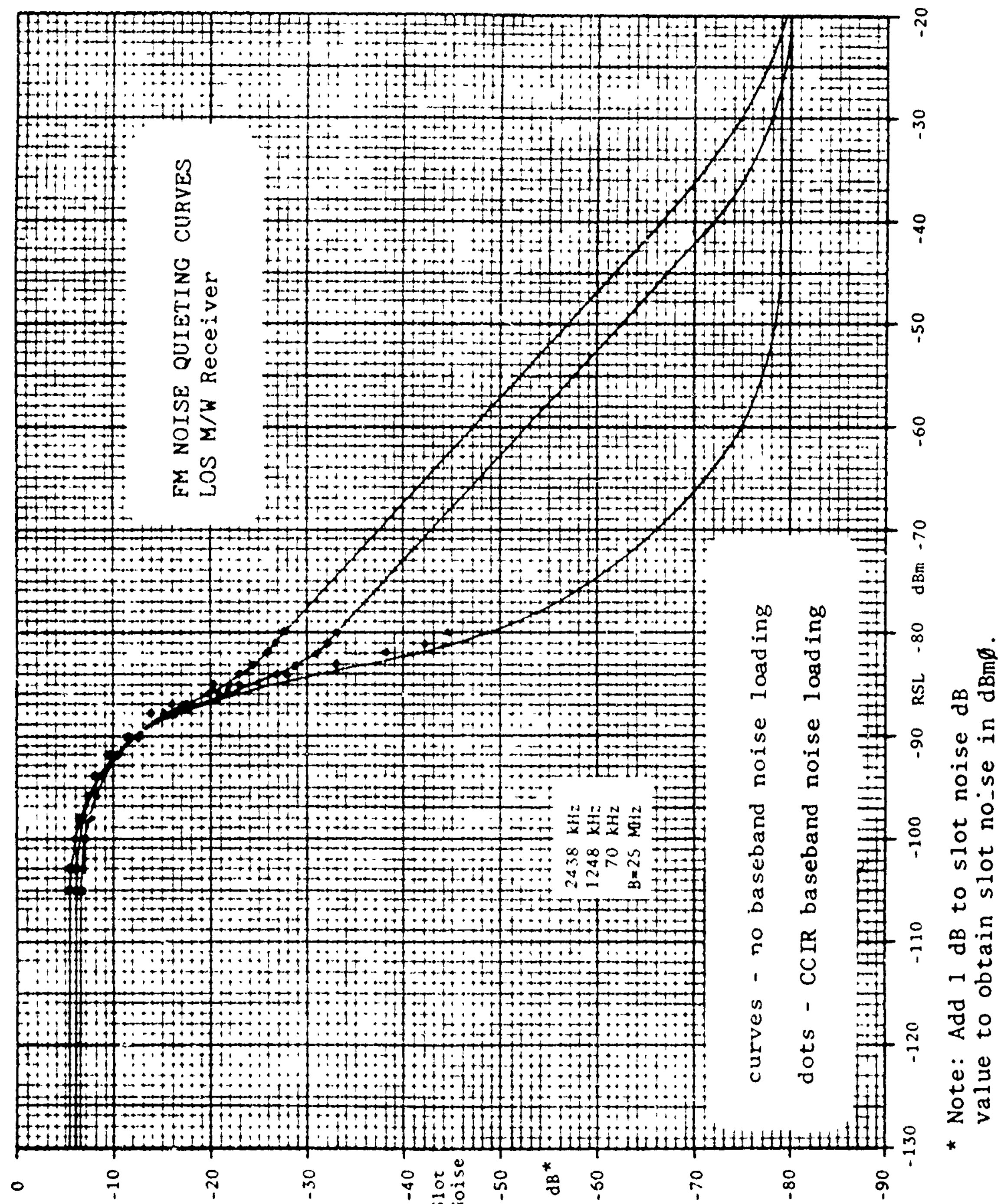
4.9 From a theoretical point of view, it has been shown by many sources that as FM threshold is approached, the baseband noise becomes impulsive in character. It is not at all obvious whether or not this would cause a change in the impulse noise in a narrow baseband slot. In other words, does the impulse noise in an FDM channel become higher (relative to the idle channel noise) as FM threshold is approached. This question was approached experimentally by using a frequency selective voltmeter to convert the baseband slot noise to a 3.1 kHz channel. To avoid clipping the noise peaks, the FSV amplifiers were operated at a very low level. However, the demodulated baseband noise was always held well above the residual noise in the frequency selective voltmeter The results were spot checked running the frequency selective voltmeter amplifiers with 10dB more gain. The results were also checked using a differnt type of frequency selective voltmeter as well as an actual frequency division multiplexer (AN/UCC-4). The results were essentially the same in all cases. The impulse noise was measured using a conventional impulse noise measuring set. Demodulated slot noise was measured using a noise measuring set with no noise weighting. Both devices had 1dB bandwidths wider than 5 kHz. Therefore, the noise measurement bandwidth was always the 3.1 kHz of the FSV.

4.10 The peak value of an impulse is difficult to define; it is theoretically infinite. Impulse noise is characterized by excursions of the total noise instantaneous voltage waveform which are much greater than the rms voltage (average power) level of the noise. The definition of peak level was defined for purposes of this report as the minimum dB voltage level above the average power level in the 3.1 kHz noise slot (for the given C/N) which would yield no indication on the impulse level measuring set when observed for a period of thirty seconds. The choice of peak level was arbitrary. If the level had been chosen as the minimum level which would cause no more than one level indication every 5 seconds, the peak level measured would have been about 2 dB lower than the level measured using the above definition. The level indications were essentially continuous at a level 4 dB lower than the level measured using the above definition. Reference levels on the

C/N (dB)	Theoretical f/B 0.0	0.0028	0.05	0.1	Measured f/B 0.0028	0.05	0.1
$-\infty$	0	0	0	0	0	0	0
0	0.2	0.2	0.2	0.2	0	1	0
+2	0.2	0.2	0.2	0.2	1	1	0
+4	0.4	0.4	0.4	0.3	2	1	1
+5	0.5	0.5	0.5	0.4	1	1	1
+6	0.6	0.6	0.5	0.4	1	1	0
+7	0.7	0.7	0.5	0.3	3	2	1
+8	0.8	0.8	0.4	0.1	3	0	0
+9	1.0	1.0	0.2	0	3	0	0
+10	1.2	1.1	0	0	3	0	0
+11	1.4	0.7	0	0	3	0	0
+12	1.7	0.1	0	0	4	0	0
+13	1.9	0	0	0	-	-	-
+14	2.3	0	0	0	-	-	-
+15	2.6	0	0	0	-	-	-

Baseband Slot Noise Increase (dB)
due to CCIR Gaussian White Noise
Baseband Loading

Table 16



FM NOISE QUIETING CURVES
LOS M/W Receiver
curves - no baseband noise loading
dots - CCIR baseband noise loading
2438 kHz
1248 kHz
70 kHz
B=25 MHz
-130 -120 -110 -100 -90 RSL -80 dBm -70 -60 -50 -40 -30 -20
0 -10 -20 -30 -40 -50 -60 -70 -80 -90
Slot Noise
dB*
* Note: Add 1 dB to slot noise dB
value to obtain slot noise in dBmØ.

Figure 29

impulse measuring equipment had a minimum step size of 2 dB. This led to some granularity in the measured results. The experimental results indicate that impulse noise peak to rms voltage levels in narrow slots are completely independent of RSL or C/N.

4.11 It will be noticed that the impulse noise peaks were higher in the lowest frequency slots than in the others. After taking the measurements, the frequency stability of the generator used to simulate the RSL was checked on a frequency counter. The signal was observed to jump around the nominal center frequency as much as 50 KHz. The signal generator was replaced by the actual microwave transmitter. The actual microwave transmitter was frequency stabilized and showed no frequency jitter at all. The results were checked in the two low frequency baseband slots. Under these conditions the results were identical to those of the other slots. It is believed that the unusually high peak noise readings in the low slots were caused by the frequency jitter of the CW generator producing occasional high level low frequency noise spikes (beat products) in the radio baseband.

4.12 The next experiment was to investigate how closely a quieting curve taken with a typical continuous wave (CW) M/W signal generator compares with a quieting curve taken with an actual M/W transmitter. The following page shows the comparison. The M/W transmitter used to make the comparison had a low power (½ watt) output and a high power (5 watts) output. The high power output was just the low power output amplified by a Traveling Wave Tube (TWT) amplifier. The quieting curve was taken separately using each output as the quieting source. There was no difference in the quieting curves. This was interesting (but predictable) since TWTs operating in the frequency range of the LOS receiver (5 gHz) have noise figures of typically 25 dB. When the quieting curves are compared, it will be noticed that the only significant difference is in the high C/N region. This is indicative of the frequency noise characteristics of the two quieting sources.

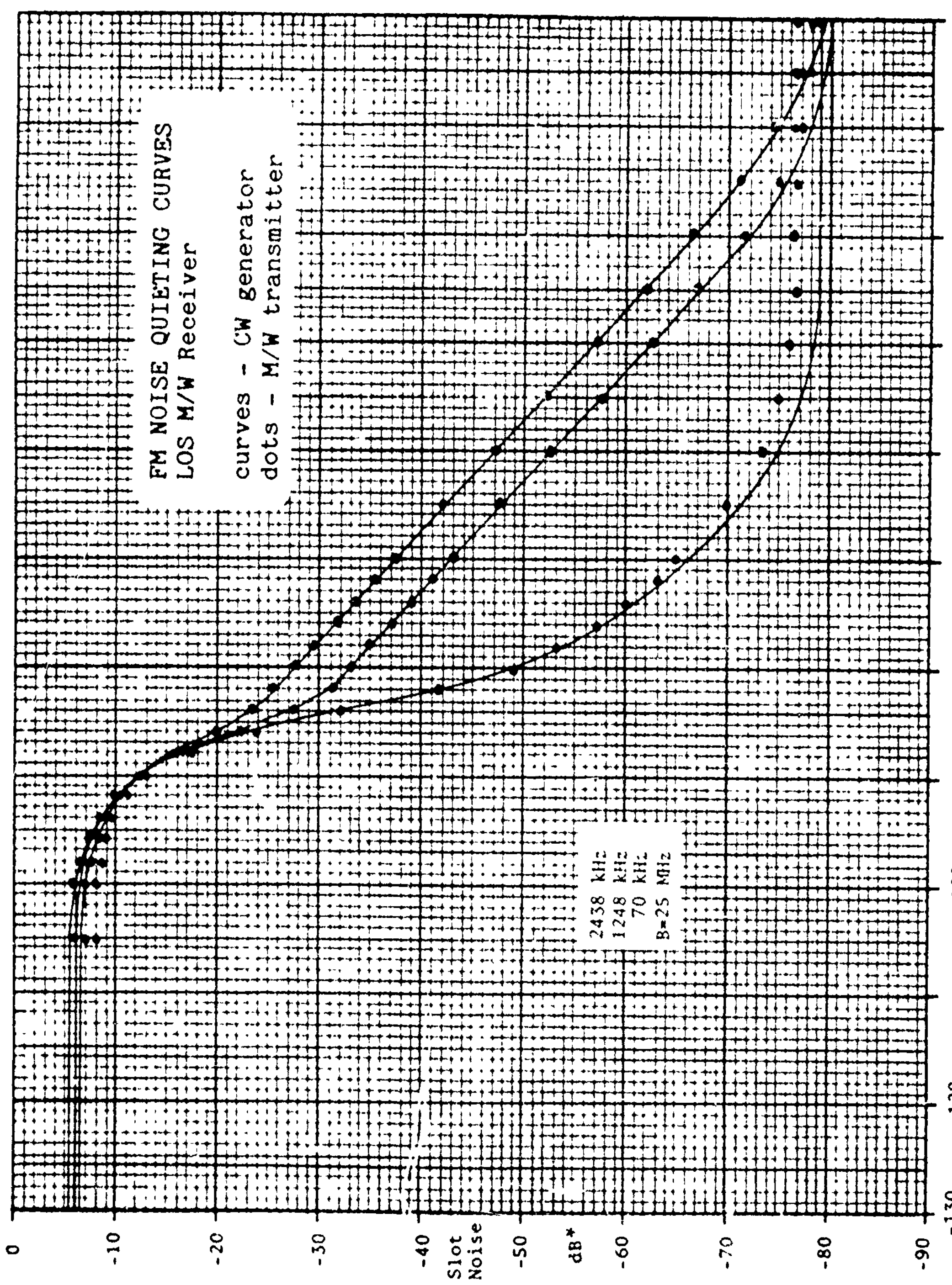
4.13 Sweeping generators (even in the CW mode) are notorious for their poor phase/frequency noise performance. Quieting curves were taken using two different sweeping microwave oscillators (data courtesy of H.L. Bennetts). Notice the wide range of possible values for slot noise in the high C/N region when different sources are used. Sweeping generator B (the up convertor for the HP microwave link analyzer) gave the best performance of any of the generators tested. In the high C/N region, its noise performance was as much as 5 dB better than the noise performance for comparable conditions with the actual microwave transmitter. The quieting curves illustrate the futility in using a quieting signal generator of unknown frequency noise performance to evaluate the high C/N slot noise performance of a M/W receiver. The inherent noise of the receiver is sometimes less than the noise of the quieting source.

4.14 Next, quieting curves were taken using wide slots. The two relatively narrow bandwidth slots were formed using NPR test set filters. The

Table 17

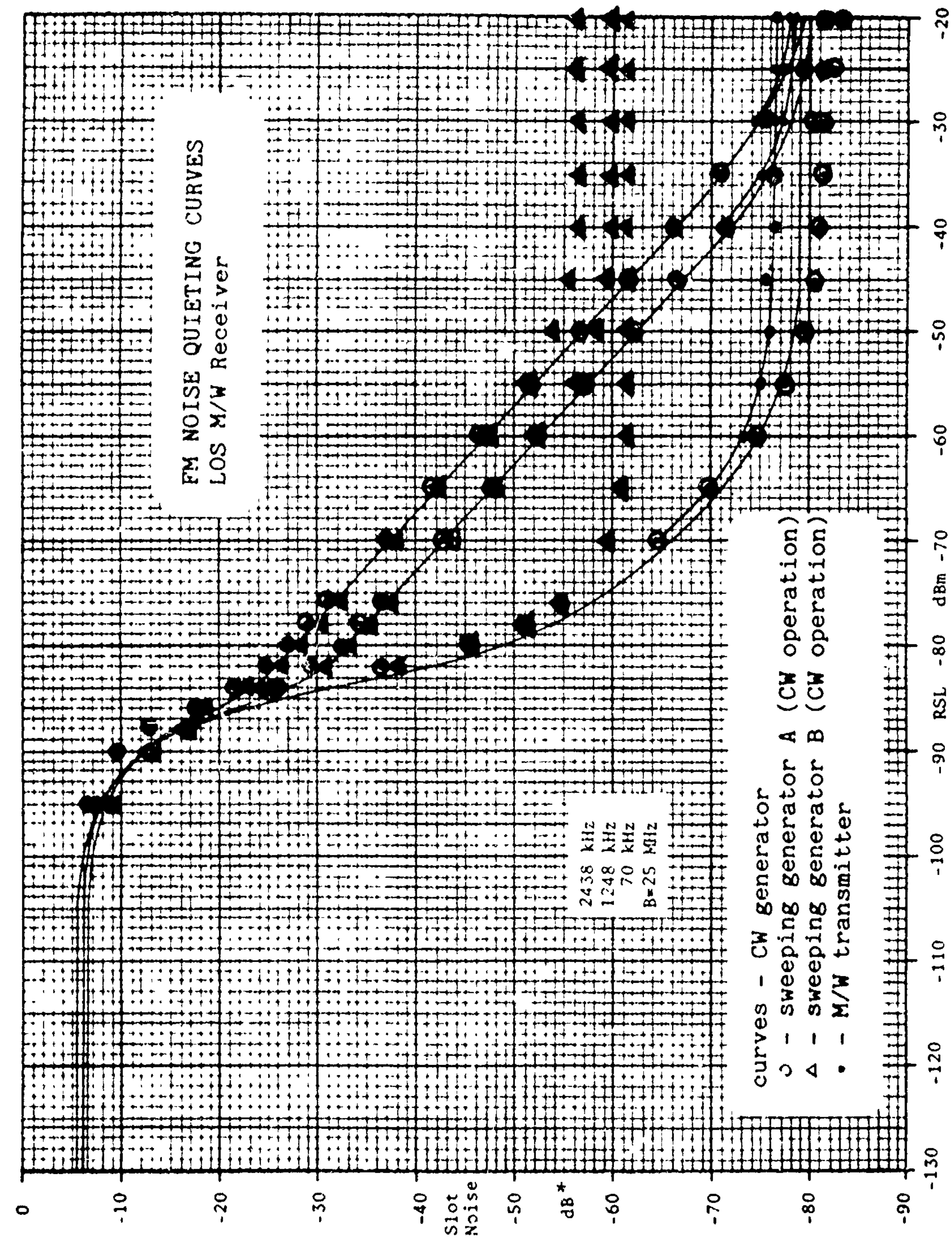
RSL (dBm)	f/B (f) 0.1 (2.5 MHz)	0.05 (1.25 MHz)	0.025 (625 KHz)	0.01 (250 kHz)	0.0025 (62.5 kHz)	0.001 (25 kHz)
-120	16.5	16	16	16	16	16
-100	14.5	16.5	16.5	16	16.5	16.5
- 90	15	15.5	15.5	16	15.5	15.5
- 80	16	16	15	15	16.5	16
- 70	16.5	15.5	15.5	15.5	15.5	19
- 60	16	15.5	17.5	15.5	14.5	19.5
- 50	16	15.5	15.5	16	17	19
- 40	16	15	15.5	16.5	17.5	20
- 30	16.5	16	15.5	15.5	17.5	25.5
- 20	16.5	16	15.5	15.5	18	23 5

Impulse Noise Peak (dB) in a 3 kHz Baseband Slot Versus RSL
(0 dB noise reference is ICN in noise sl ,t at given RSL)



* See note, Figure 29.

Figure 30



* See note, Figure 29.

Figure 31

wide bandwidth slot was formed using the baseband low pass filter in the M/W receiver itself. The noise was measured using a wide bandwidth true rms meter. The various slot measurements are plotted by themselves and also versus the normal narrow slot quieting curves. For comparison with the quieting curve data, 10 log (filter bandwidth (kHz) / 3.1kHz) was subtracted from the wide slot data. FM threshold can be determined for the wide slots by integrating the narrow slot noise formula over the frequency range of the wide slot. As noted earlier, this leads to a result which indicates an FM threshold for the widest slot should occur about one dB earlier than FM threshold for the 2438 kHz narrow slot. The experimental evidence shows reasonable agreement with that result.

4.15 After observing the noise characteristics of the M/W receiver, the next set of experiments observed the baseband signal suppression characteristics of the M/W receiver. The modulated RF signal was then applied to the LOS M/W receiver through a variable attenuator. The test tone suppression as a function of C/N (RSL) was observed using a frequency selective voltmeter (FSV) at the baseband of the receiver. Since the noise power in the FSV slot was significant at low C/N ratios, the actual FSV measurement was test tone plus noise power rather than just test tone power. To correct for this, a FSV measurement was taken with the test tone applied to the transmit baseband. Another measurement (of noise) was made with the test tone removed from the baseband. By subtracting the noise power (in milliwatts) from the test tone plus noise (in milliwatts) the FSV measurements were corrected for the effects of noise. The test tone levels used were less than normal CCIR loading to avoid introducing significant increases in receiver slot noise due to the modulation of the received signal.

4.16 The first experiment was to observe signal suppression at three baseband frequencies. The three frequencies were 0.0028, 0.05, and 0.1 times the IF bandwidth. The 0.05 frequency had the least suppression of any of the baseband test tone frequencies. The performance of the other test tones relative to the performance of the 0.05 test tone has been tabulated on the next page.

4.17 The next experiment was to observe signal suppression as a function of baseband loading. Using an NPR test set white noise generator, CCIR recommended values of white noise loading were applied to the transmitter baseband to simulate normal baseband loading. Using the NPR test set bandstop filters, noise was suppressed at three different frequencies. Test **tones** were placed in the transmitter baseband at those frequencies. The test tone suppression was measured with no white noise loading and was then measured with full noise loading. The difference in the baseband suppression for the two conditions **is shown in Table 18.**

4.18 Next, the baseband test tone suppression values obtained during the previous experiments is plotted on an expanded vertical scale.

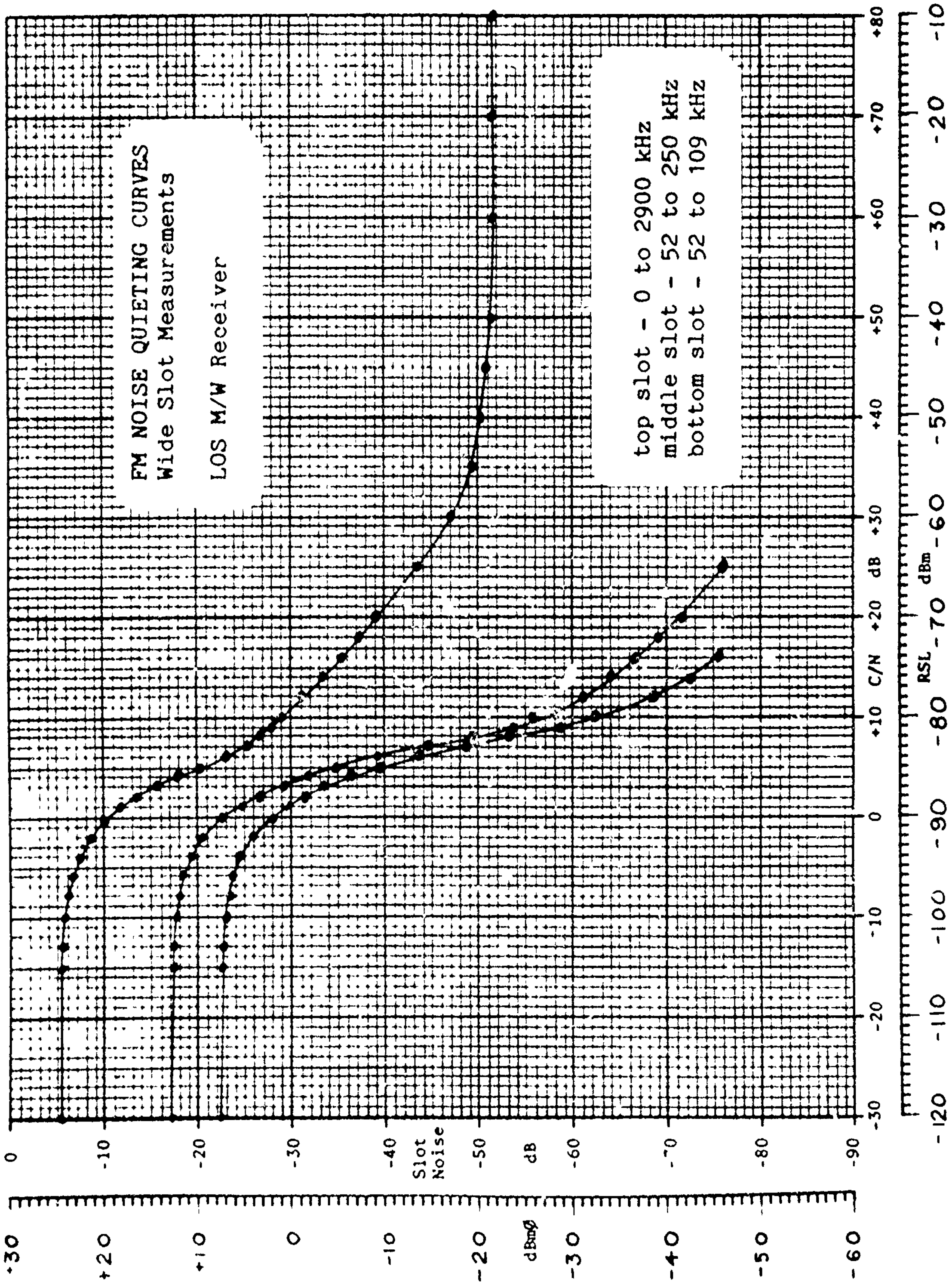
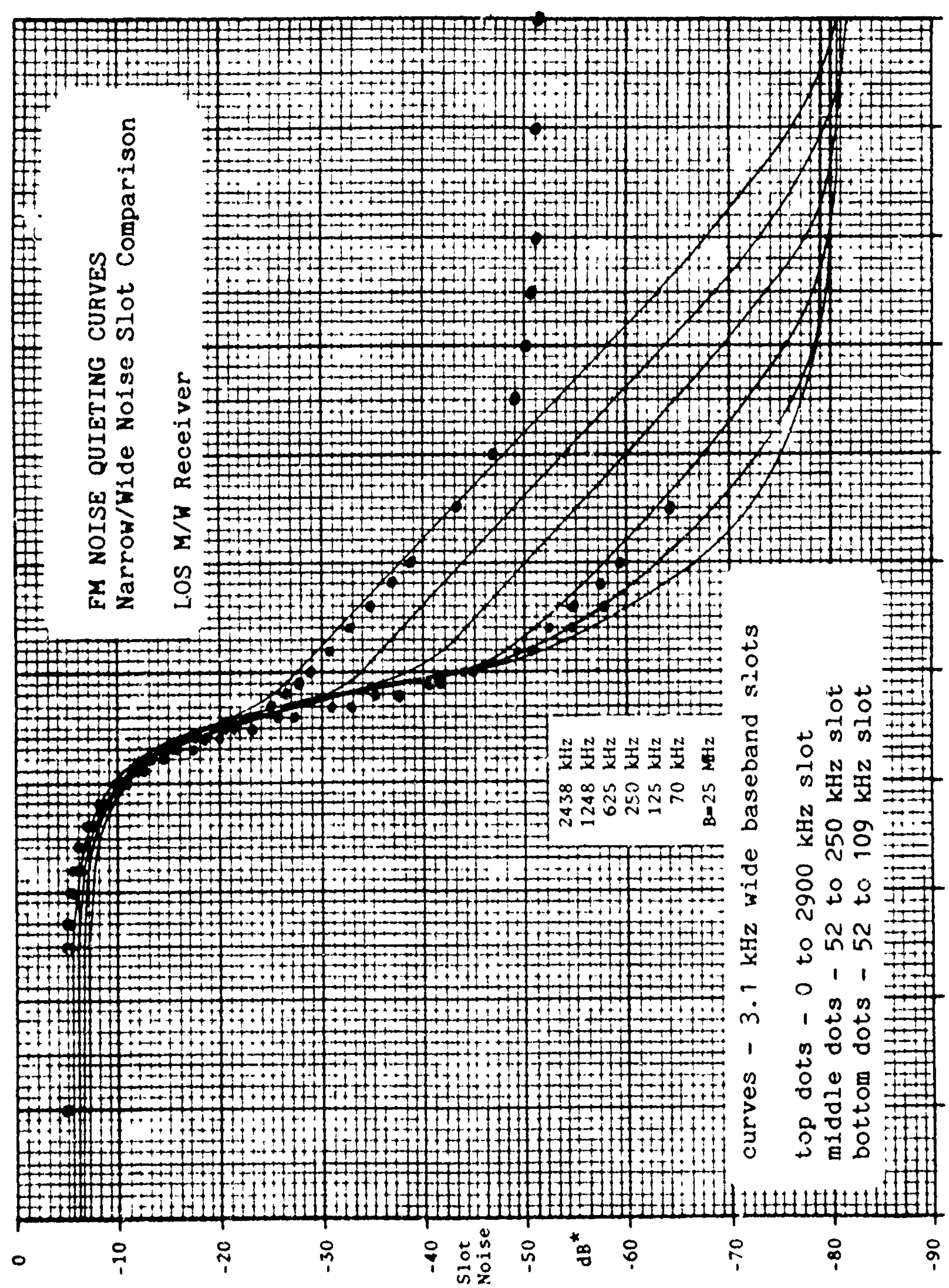

FM NOISE QUIETING CURVES
Wide Slot Measurements
LOS M/W Receiver
top slot - 0 to 2900 kHz
middle slot - 52 to 250 kHz
bottom slot - 52 to 109 kHz
C/N dB
RSL dBm
Slot Noise dB
dBmØ

Figure 32



* See note, Figure 28.

Figure 33

C/N (dB)	f/B	
	0.0028	0.1
-3	-0.5	-0.6
-2	-0.9	-0.7
-1	-0.5	-0.9
0	-0.8	-1.0
+1	0.0	-0.6
+2	-0.8	-0.8
+3	-0.3	-0.6

Baseband Signal Suppression (dB) at High and Low Baseband Frequencies Relative to Signal Suppression at the Middle of the Baseband (reference test tone at normalized baseband frequency of f/B = 0.05)

C/N (dB)		f/B	
	0.0028	0.05	0.1
-3	0.0	0.0	0.0
-2	0.0	0.0	0.0
-1	0.0	-0.5	0.0
0	0.0	-0.4	0.0
+1	0.0	-1.0	0.0
+2	0.0	0.0	0.0
+3	0.0	0.0	0.0

Baseband Signal Suppression (dB) with CCIR Gaussian White Noise Baseband Loading Relative to Signal Suppression with no Baseband Noise Loading

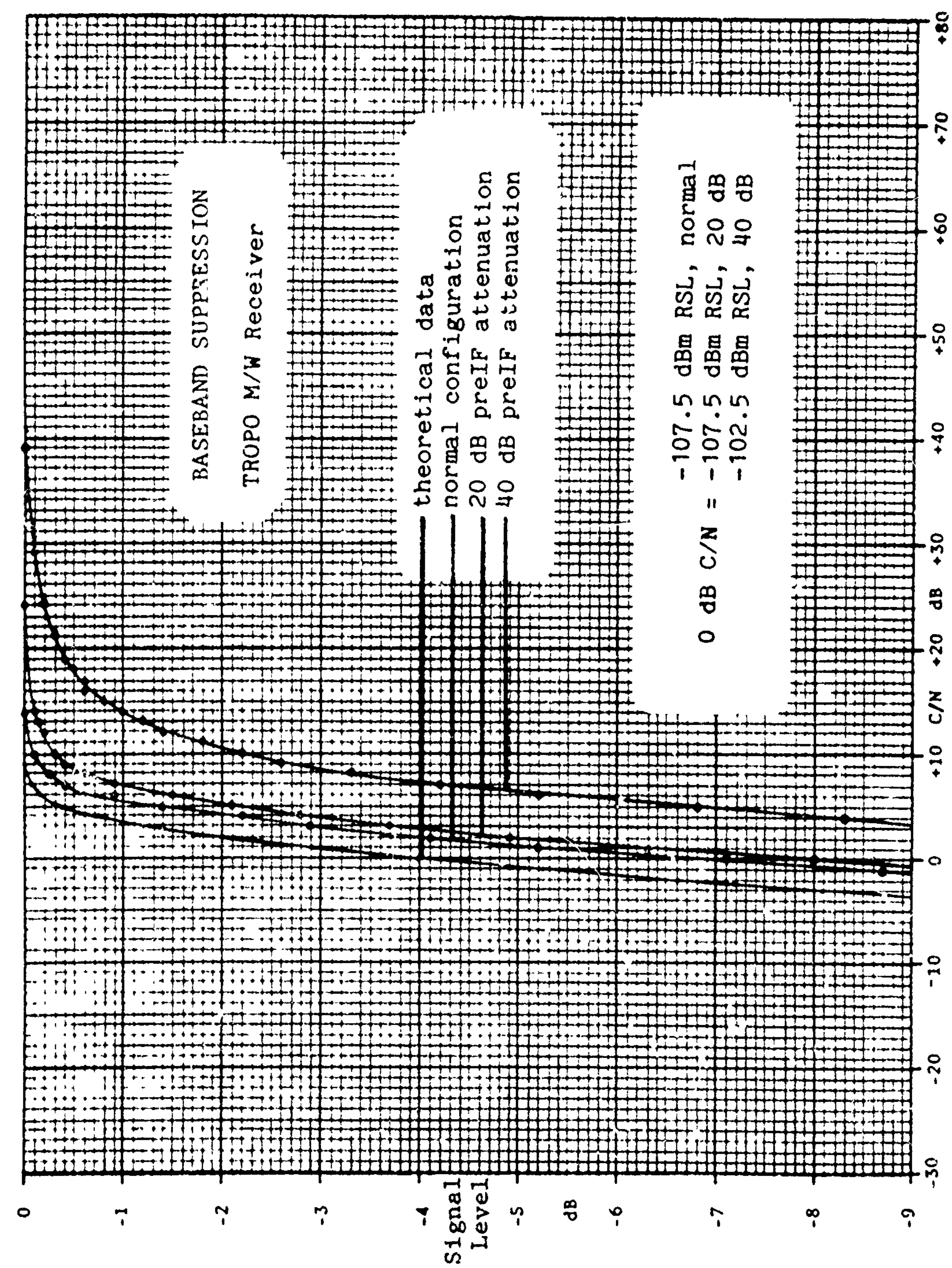
Table 18

The graphs on the next two pages show the effect of changes in limiter action. The C/N values have been corrected to compensate for changes in receiver noise figure due to the introduction of preIF attenuation. Note that as limiter action goes soft, baseband suppression starts at earlier C/N values. Also, notice that the LOS receiver is much more sensitive to preIF signal loss than is the TROPO receiver.

4.19 In elementary developments of the theoretical noise of FM receivers, one of the assumptions is that the received signal is centered in the passband of the receiver. The frequency response of the passband is also assumed to be symmetric about the carrier frequency. The unmodulated receive carrier signal was deliberately tuned to one side of the receiver's IF response and quieting curves were accomplished. Data was taken in the 2438, 1248, and 70 kHz slots and is graphed on the next page. The data reflects an extreme violation of the previous assumptions. It can be anticipated that receiver idle noise performance will be degraded if the received signal is not centered in the receiver's (IF) passband response or if the receiver's (IF) passband frequency response is unsymmetric relative to center frequency.

4.20 Most theoretical developments of receiver signal and noise performance assume that the noise in the receiver is strictly thermal. The noise power is assumed to be spread equally across the bandpass response of the receiver. This is not always the case in practice. The most elementary form of nonthermal noise is sine wave interference.

4.21 The next few graphs show the signal and noise characteristics of an FM M/W receiver for different levels and frequencies of unmodulated carrier interference. The baseband test tone signal frequency was 0.05 times the receiver IF bandwidth. The tests were run by mixing the outputs of the unmodulated M/W transmitter and a CW signal generator. The composite signal was applied to the input of the M/W receiver. Separate attenuators allowed the outputs of the two signal sources to be varied independently. The output of the M/W transmitter was considered the carrier and the output of the CW generator was called the interference. The carrier source was tuned to the center frequency of the receiver. Results were obtained with the interference tuned to two different frequencies. For one set of data, the interfering signal was tuned to a frequency differing by ½ MHz from the receiver center frequency. This frequency offset was chosen so that the interfering signal would be near the carrier frequency, but none of the beat products caused by crossmodulation products (beat products) in the receive baseband (due to the two RF input signals) would fall into the noise measurement slot skirts. Also, the large increase in baseband noise when two RF carriers have essentially the same frequency, but randomly varying relative phase was avoided. The other set of data was taken with the interfering signal offset 10 mHz from the receiver center frequency. This put the interfering signal at the edge of the receiver's IF passband. Although the interfering signal with the larger frequency offset caused more baseband noise, a spectrum analyzer sweep of the radio baseband would not have shown



BASEBAND SUPPRESSION
TROPO M/W Receiver
theoretical data
normal configuration
20 dB preIF attenuation
40 dB preIF attenuation
-107.5 dBm RSL, normal
0 dB C/N = -107.5 dBm RSL, 20 dB
-102.5 dBm RSL, 40 dB
0
-1
-2
-3
-4
-5
-6
-7
-8
-9
Signal Level dB
-30
-20
-10
0
+10
+20
+30
+40
+50
+60
+70
+80
C/N dB

Figure 34

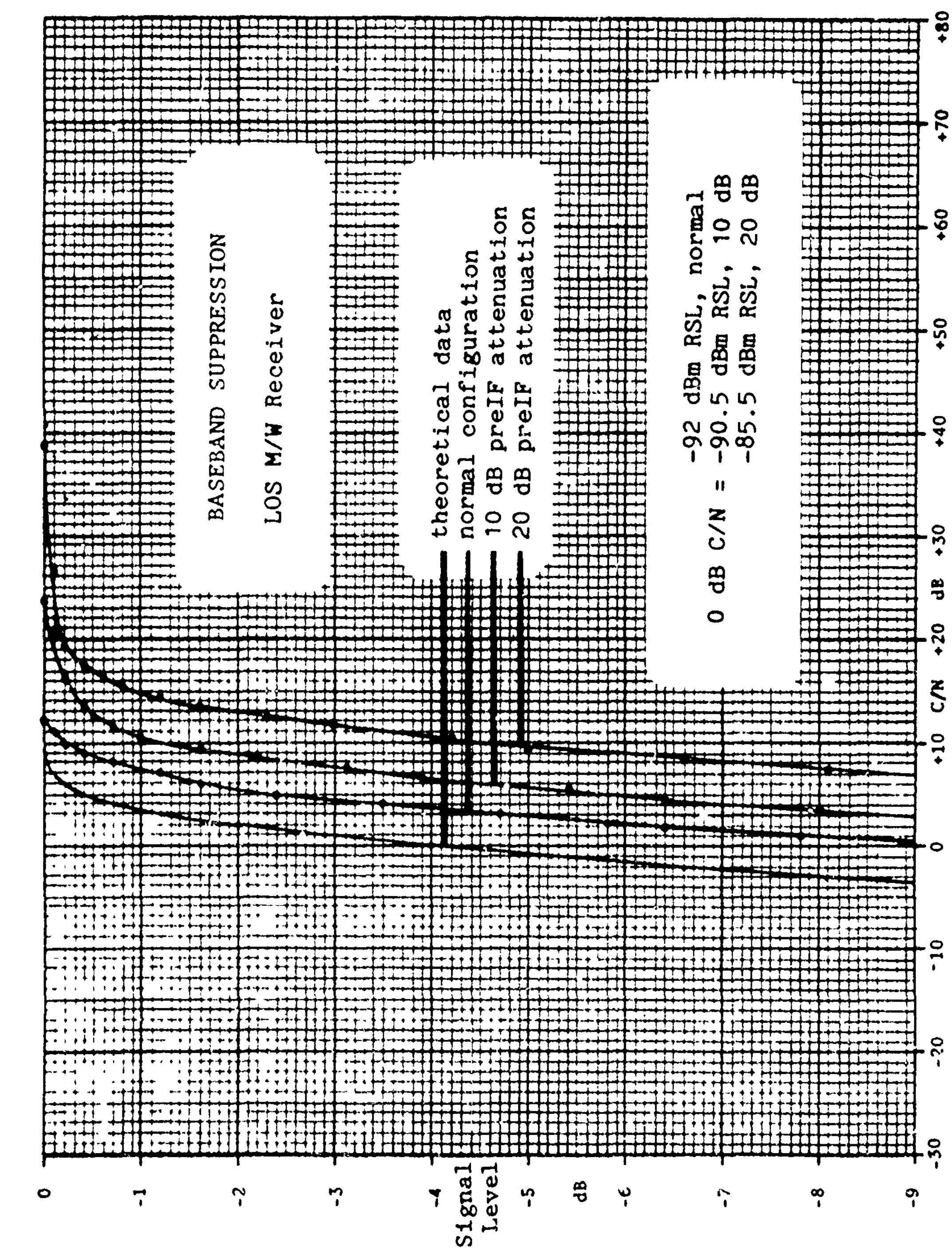
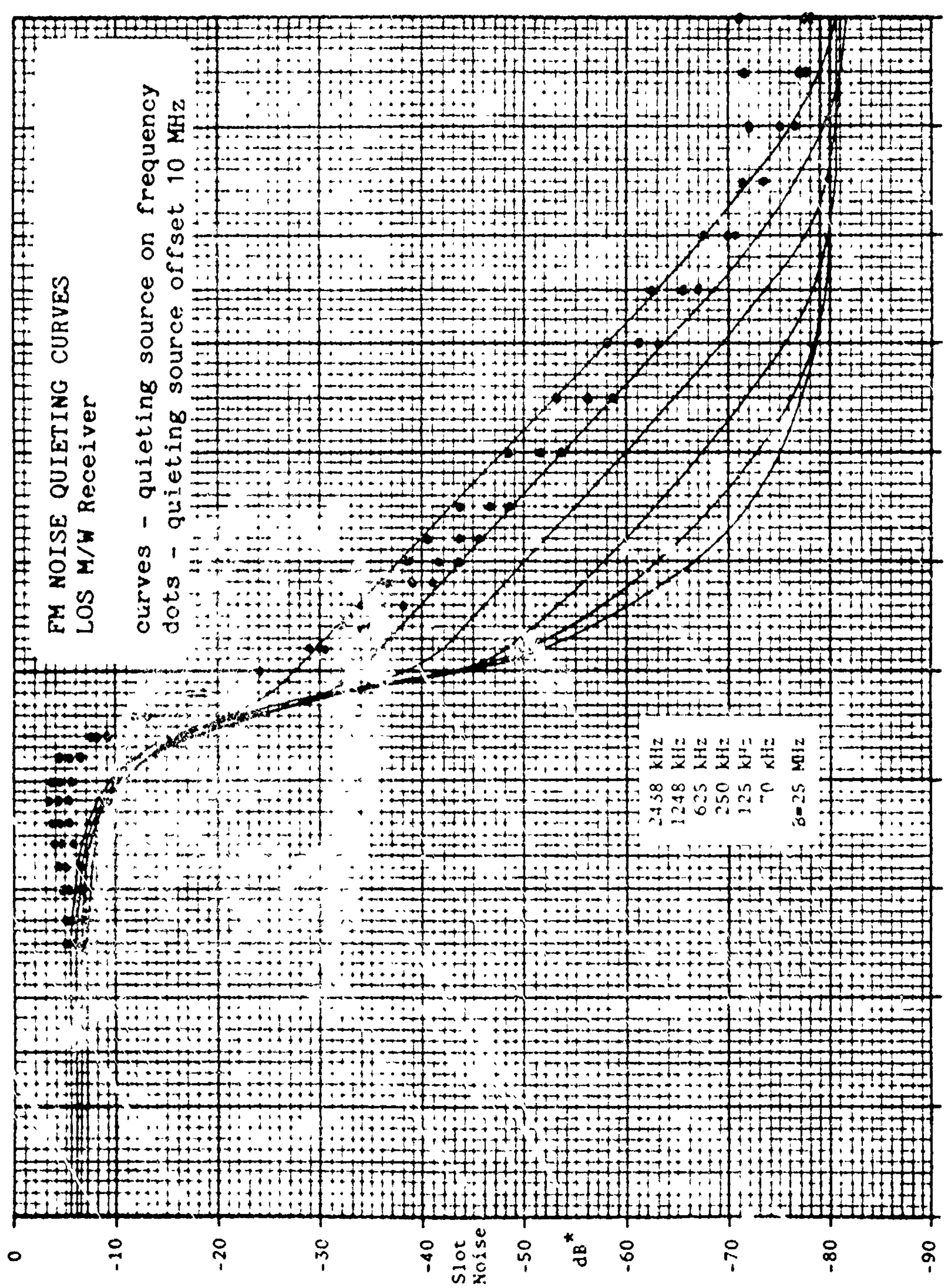

BASEBAND SUPPRESSION
LOS M/W Receiver
theoretical data
normal configuration
10 dB preIF attenuation
20 dB preIF attenuation
-92 dBm RSL, normal
0 dB C/N = -90.5 dBm RSL, 10 dB
-85.5 dBm RSL, 20 dB
Signal Level
dB
0
-1
-2
-3
-4
-5
-6
-7
-8
-9
C/N dB
-30
-20
-10
0
+10
+20
+30
+40
+50
+60
+70
+80

Figure 35



* See note, Figure 28.

Figure 36

the beat product since it would appear at a frequency much higher than the cutoff frequency of the receiver baseband low pass filter.

4.22 The first set of graphs were taken by holding the interfering signal power at -60 dBm. When the C/I ratio approaches +10 dB, the signal and noise properties become noticeably affected. Of course, if baseband slot noise had been observed at integer multiples of the frequency difference between the carrier and interference, the noise increase would have been more distinct and would have been noticed at much larger C/I ratios. The capture effect of the receiver is dramatic as it starts to quiet on the interference rather than the carrier. The second set of graphs were taken with the interfering signal level held at -90 dBm. The third set of graphs were taken with the interference level varied with carrier level variation so that the interference was always 10 dB less than the carrier level. Notice that in all cases the slot noise was noticeably higher and baseband signal suppression earlier for signals farther away from the center of the receiver IF response.

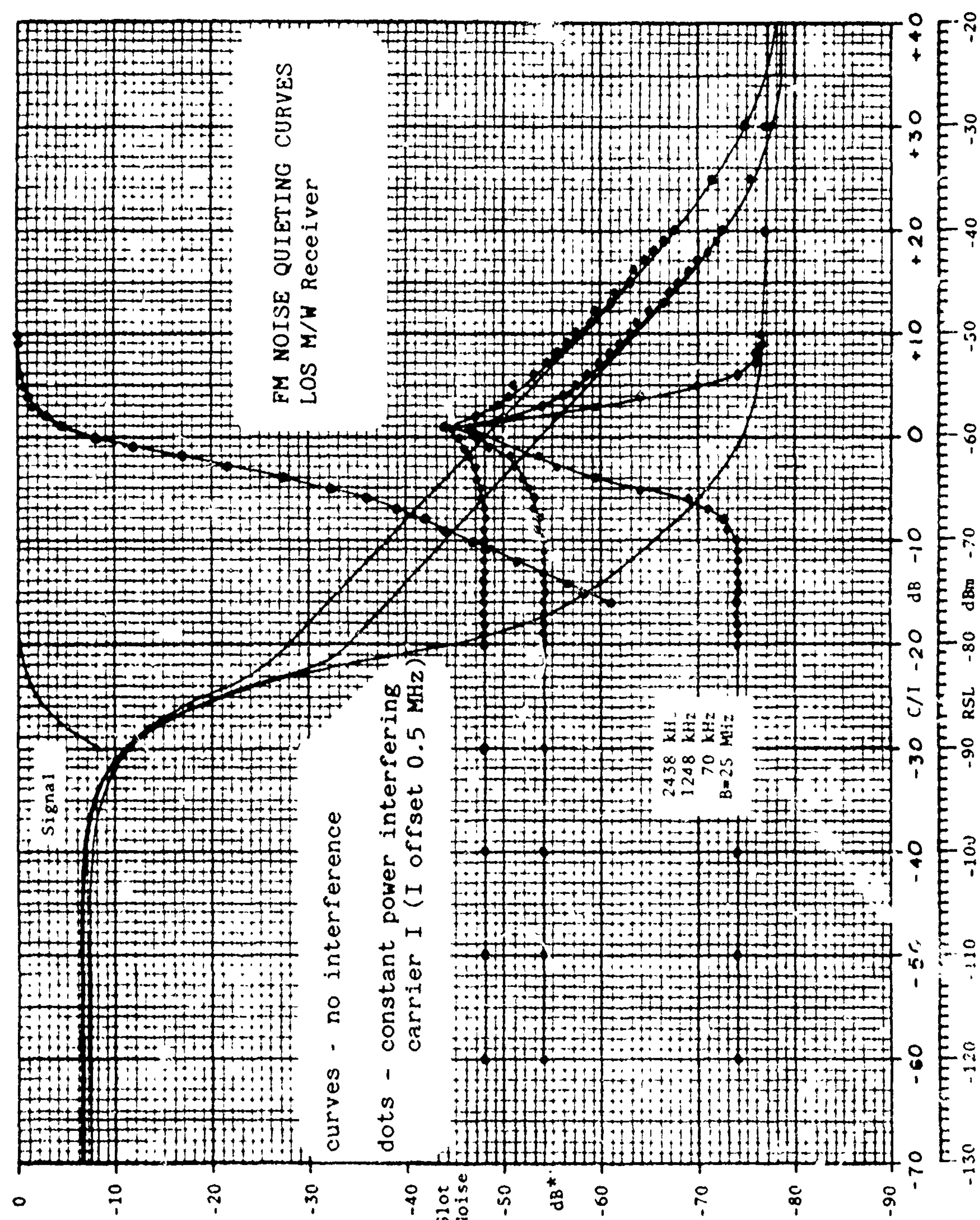
4.23 The last set of interference data was taken by holding the interfering signal power constant but varying its frequency relative to receiver center frequency. The receiver had an RF filter as well as an IF filter. The RF filter frequency response was broad enough that, for the interfering signal frequencies used for this test, the frequency selectivity characteristics of the receiver can be assumed to be entirely due to the IF frequency response of the receiver. Clearly the noise process is complex and a function not only of C/I and interference frequency offset but also baseband noise slot frequency.

4.24 It should not be assumed that the performance of all FM receivers is the same. To constrast the previous results, tests were performed on another LOS M/W receiver. It's nominal equipment characteristics were quite similar to the characteristics of the LOS M/W receiver used to perform the previous experiments. A quieting curve for the receiver is given on the following page. Note the abnormally high noise in the low frequency noise slots. Also, note the kink in the curves near the -80 dBm RSL. During the experiment it was noted that the noise measurements in this region were unusually erratic. Although it is difficult to see on the quieting curve, two of the low frequency noise slots actually cross each other, a theoretical impossibility.

4.25 The next page shows spectrum analyzer photographs of the baseband noise spectrum for various RSLs. A comparison with the corresponding photos for the other LOS receiver shows considerable difference. The center photograph highlights the noise instability for RSLs near the kinks in the quieting curve. Notice the crisscrossing of the baseband spectrum. When this photo was taken, the noise measurements were so unstable that the same picture could never be reproduced. Every trace of the spectrum analyzer produced a radically different picture. This was considerably different than the case for the LOS receiver used in

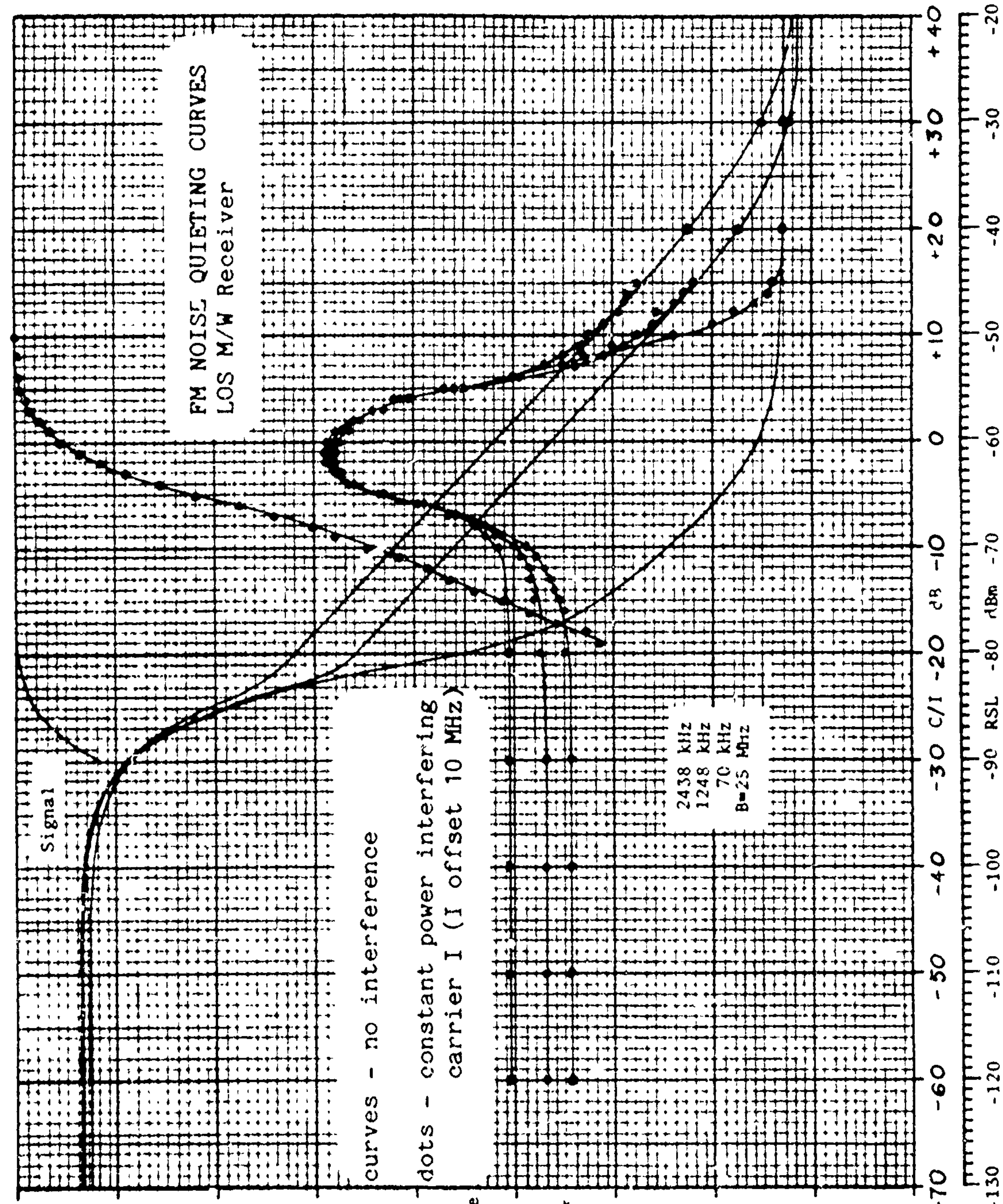
the rest of the test for this report. Results taken with the AN/FRC-157 receiver were repeatable day after day. The results taken on the other receiver changed from moment to moment. Not even the baseband suppression characteristics remained constant for the other LOS receiver. The results of tests conducted at two different times are shown on the following page.

4.26 The FM quieting curve is a sensitive indicator of noise performance of an FM M/W terminal. The following quieting curves were taken of FM M/W receivers in actual operational environments. Although it is not always clear what the problem is, the quieting curves do show that there obviously undesirable conditions at the radio terminal. Occasionally the quieting curves also shows measurement error and measurement equipment problems.



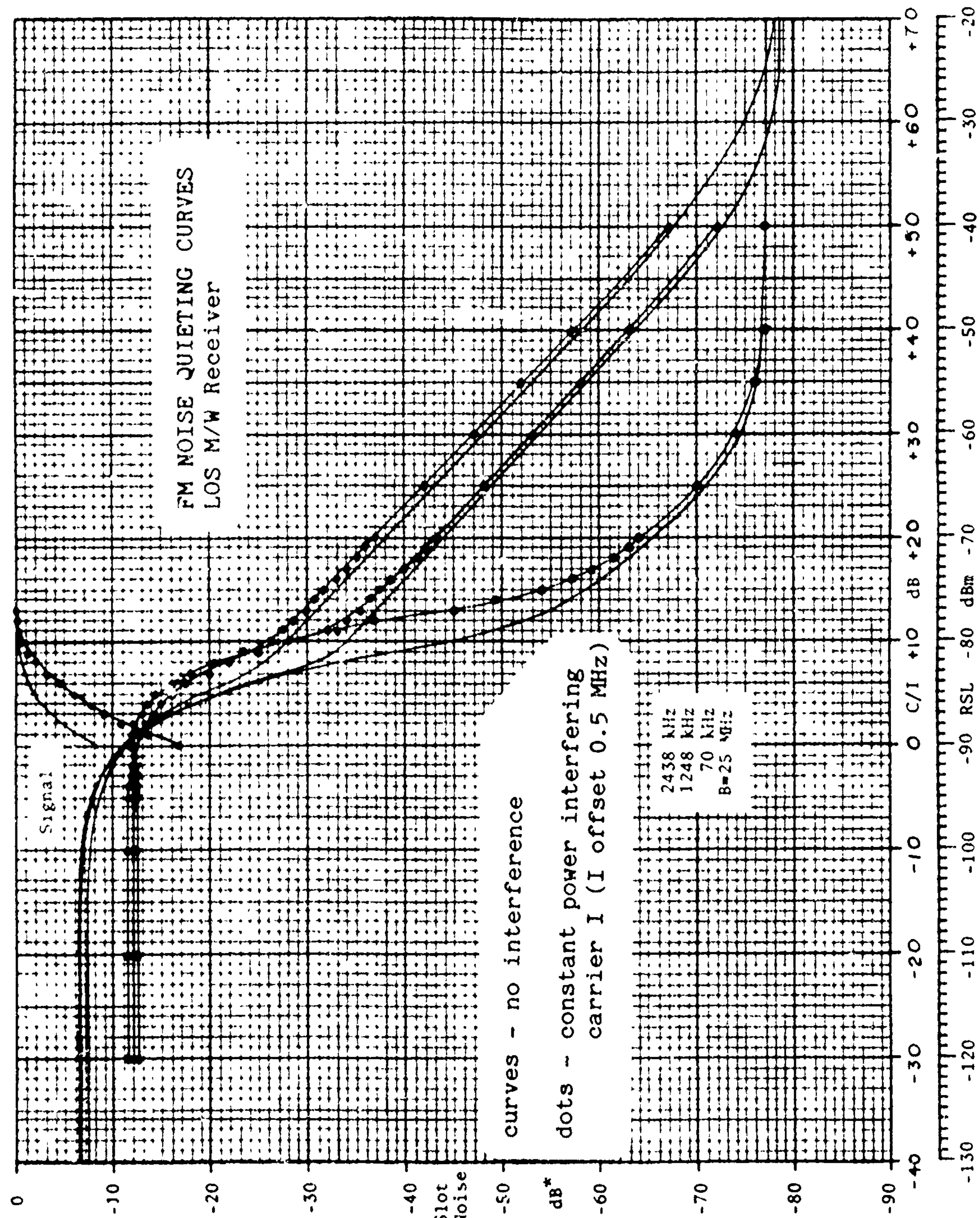
* See note, figure 28.

Figure 37



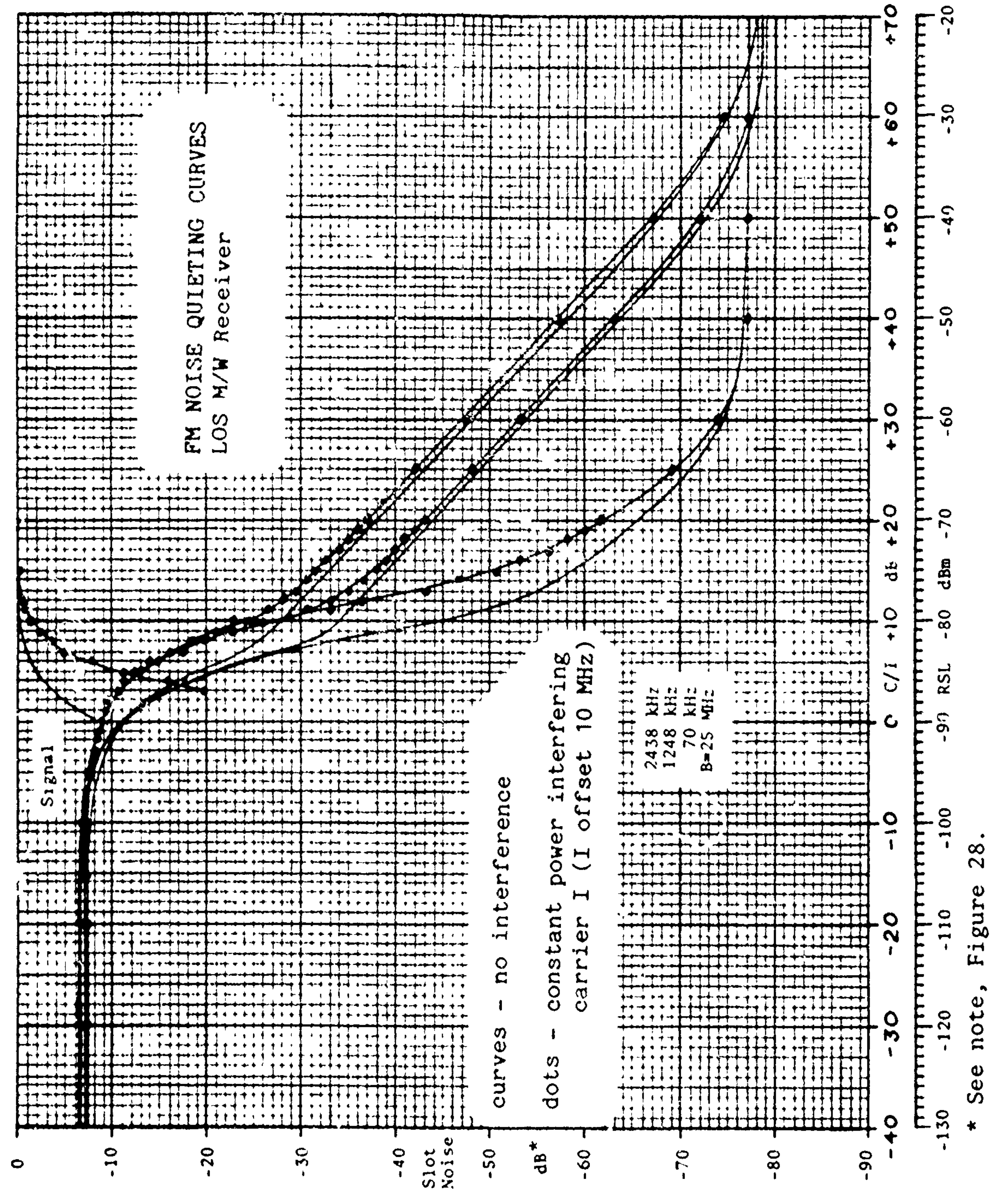
* See note, Figure 28.

Figure 38



* See note, Figure 28.

Figure 39



FM NOISE QUIETING CURVES
LOS M/W Receiver
curves - no interference
dots - constant power interfering carrier I (I offset 10 MHz)
2438 kHz
1248 kHz
70 kHz
B=25 MHz
Signal
Slot Noise
dB*
C/I dB
RSL dBm
* See note, Figure 28.

Figure 40

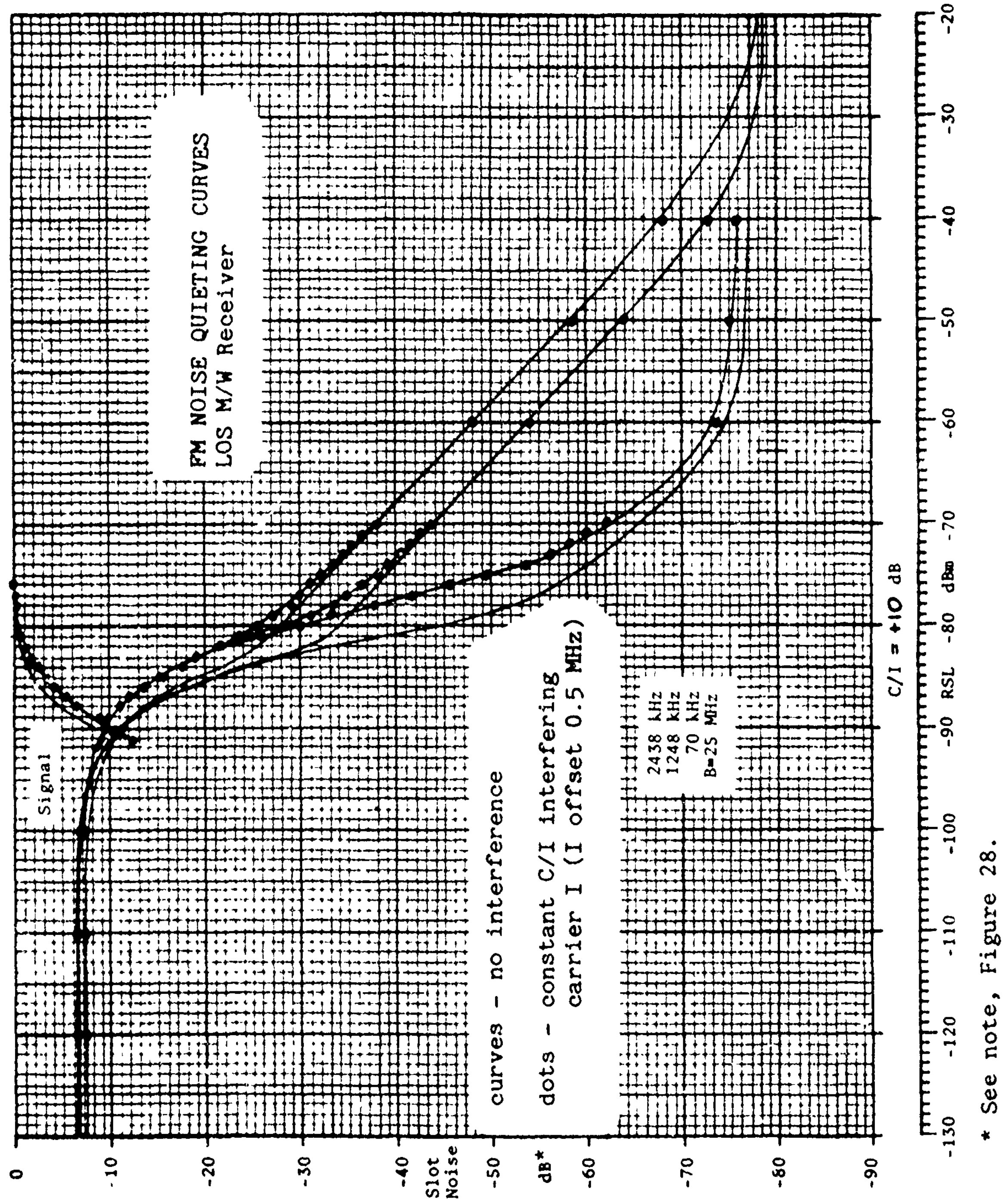
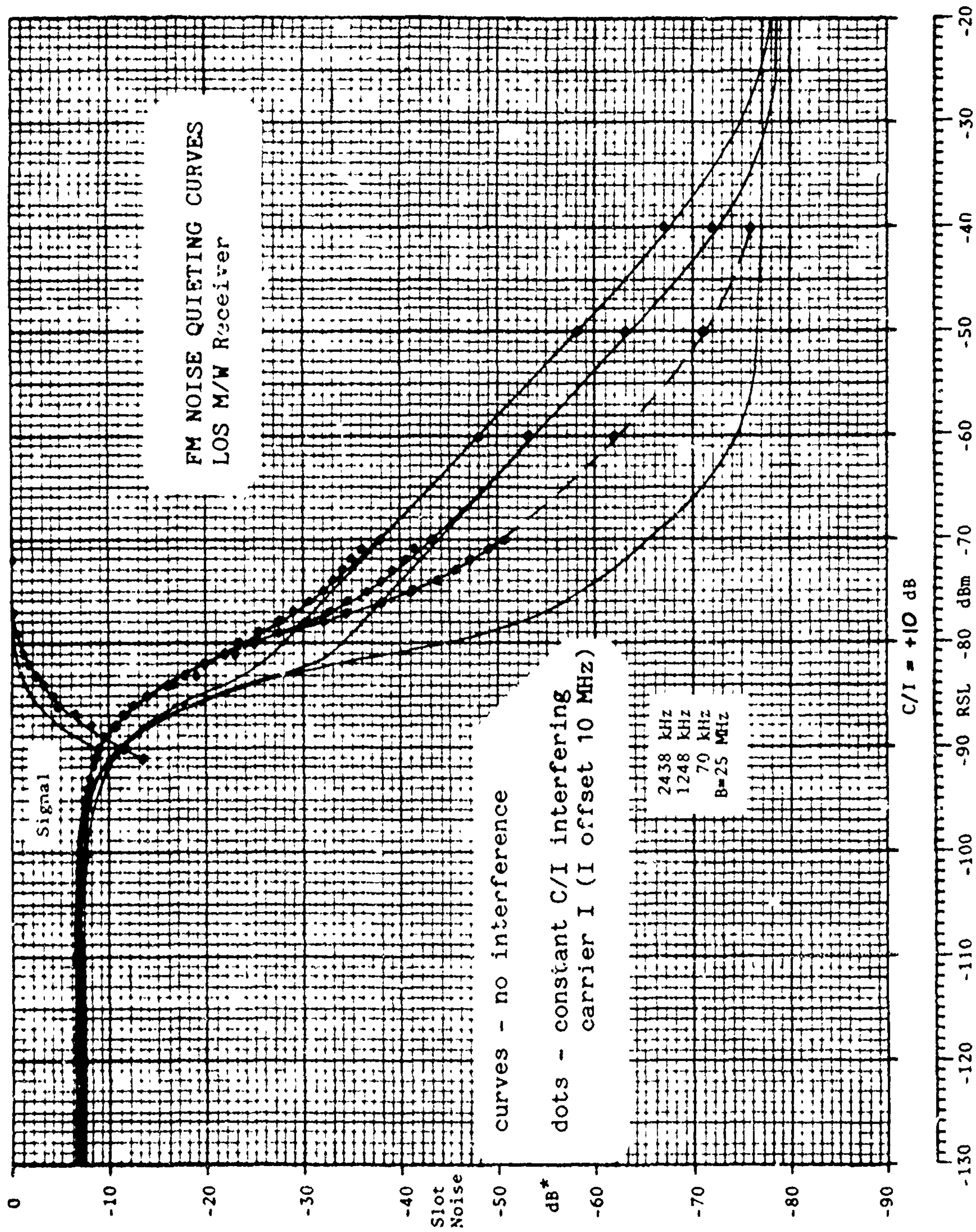
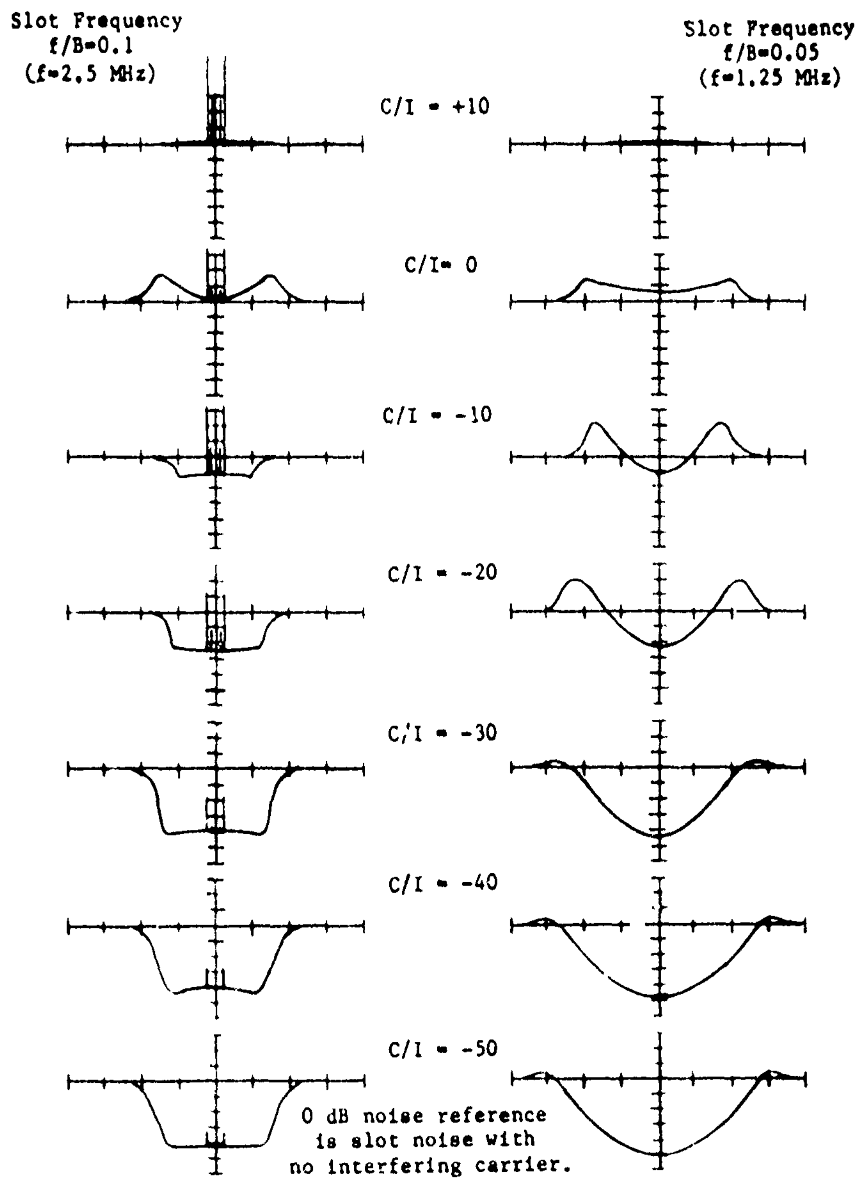

FM NOISE QUIETING CURVES
LOS M/W Receiver
curves - no interference
dots - constant C/I interfering
carrier I (I offset 0.5 MHz)
2438 kHz
1248 kHz
70 kHz
B=25 MHz
Signal
0
-10
-20
-30
-40
Slot Noise
-50
dB*
-60
-70
-80
-90
-130
-120
-110
-100
-90
RSL
-80
dBm
-70
-60
-50
-40
-30
-20
C/I = +10 dB
* See note, Figure 28.

Figure 41



* See note, Figure 28.

Figure 42



Vertical Scale : Slot Noise (10 dB/div., + above line)
Horizontal Scale : Offset (10 MHz/div.)
Intersection of lines is zero on both scales.

LOS M/W Receiver
Quieting Carrier Power (C) -60 dBm
Interfering Carrier Power (I) Variable
(Beat products neglected on right hand plots.)

Baseband Slot Noise Vs Interfering Carrier Frequency and Power

Figure 43

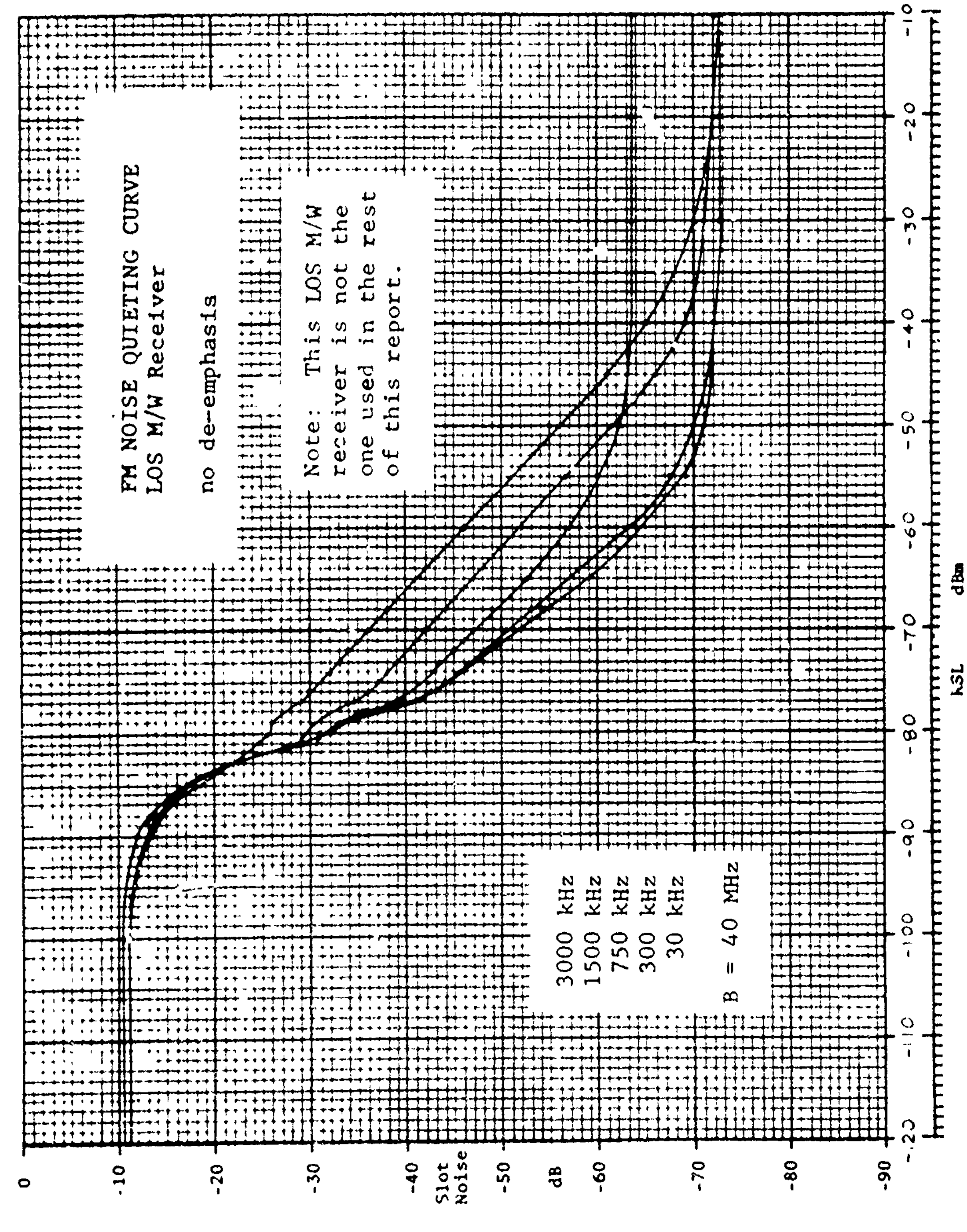
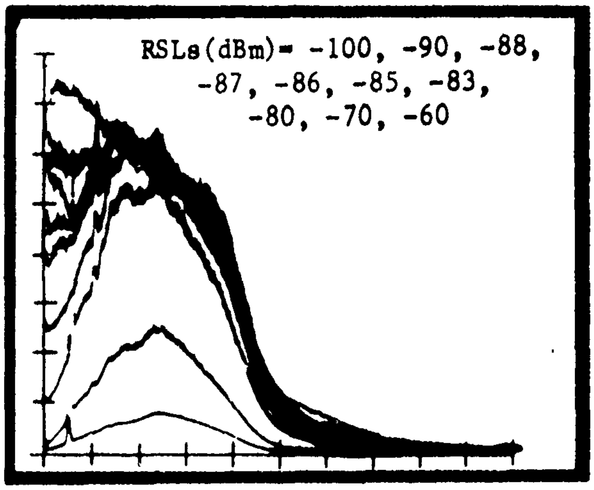
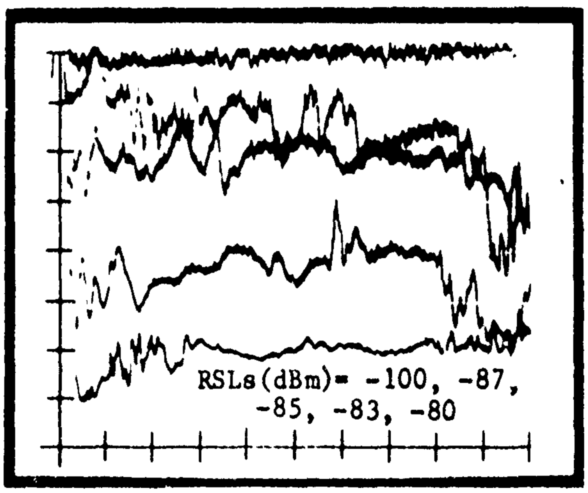

FM NOISE QUIETING CURVE
LOS M/W Receiver
no de-emphasis
Note: This LOS M/W receiver is not the one used in the rest of this report.
3000 kHz
1500 kHz
750 kHz
300 kHz
30 kHz
B = 40 MHz
Slot Noise
dB
RSL dBm

Figure 44



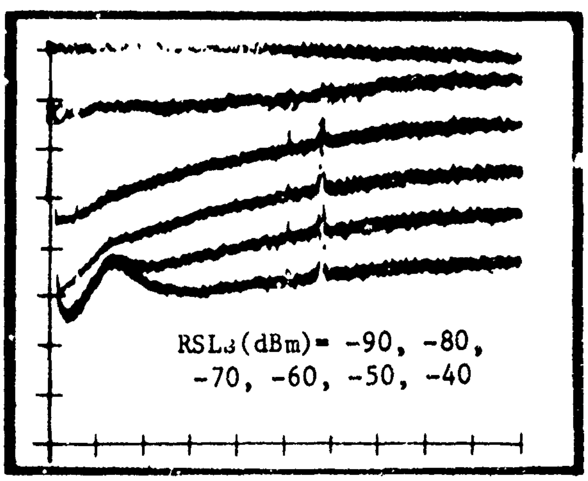
LOS M/W Receiver
no de-emphasis

Vertical: mV
Horizontal: 5 MHz/Div.
(0 to 50 MHz)



Vertical: mV
Horizontal: 200 kHz/Div.
(0 to 2 MHz)

Note: This LOS M/W receiver is not the one used in the rest of this report.



Vertical: mV
Horizontal: 500 kHz/Div.
(0 to 5 MHz)

Baseband Noise Spectrum
for Various RSLs

Figure 45

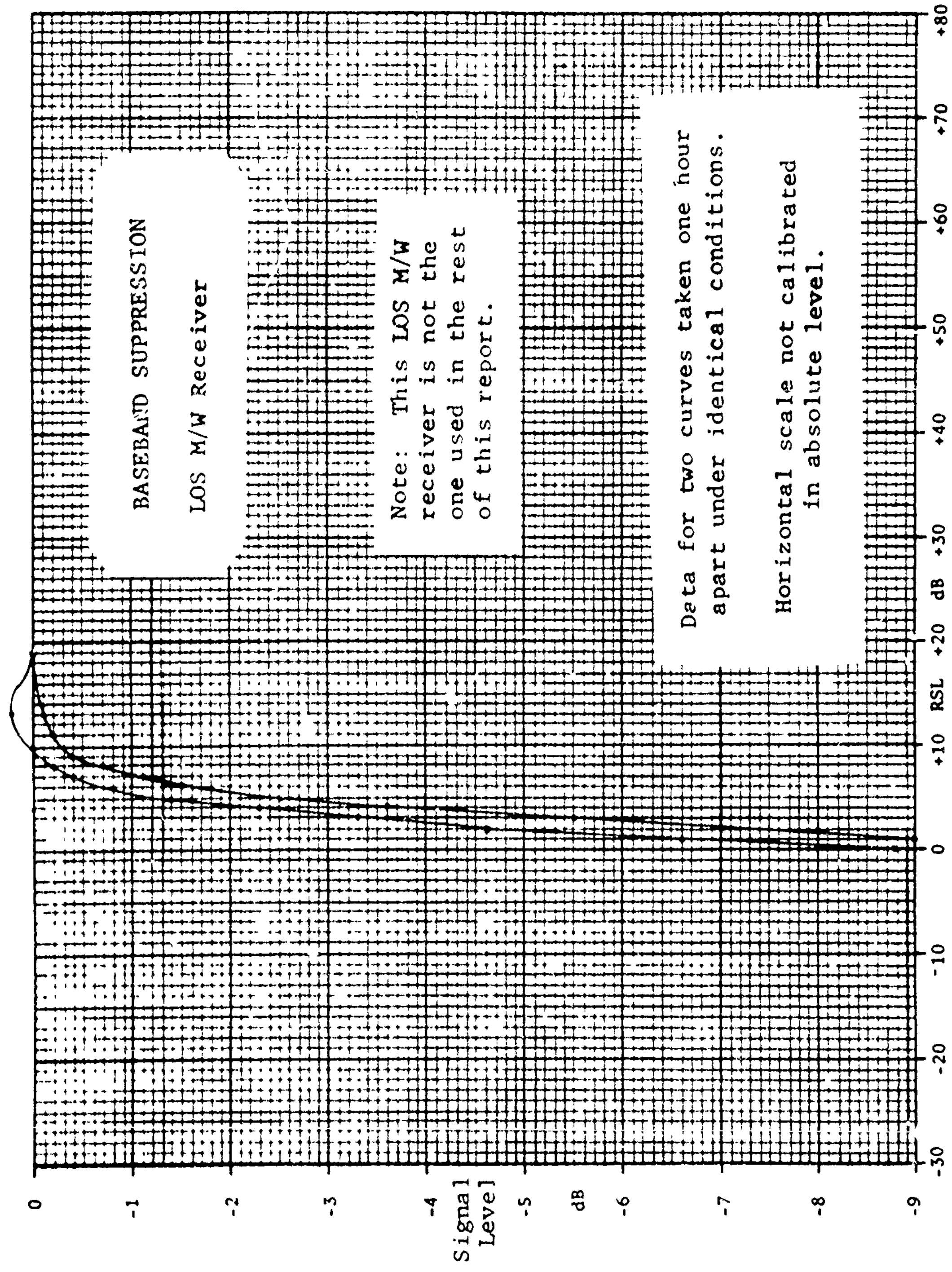
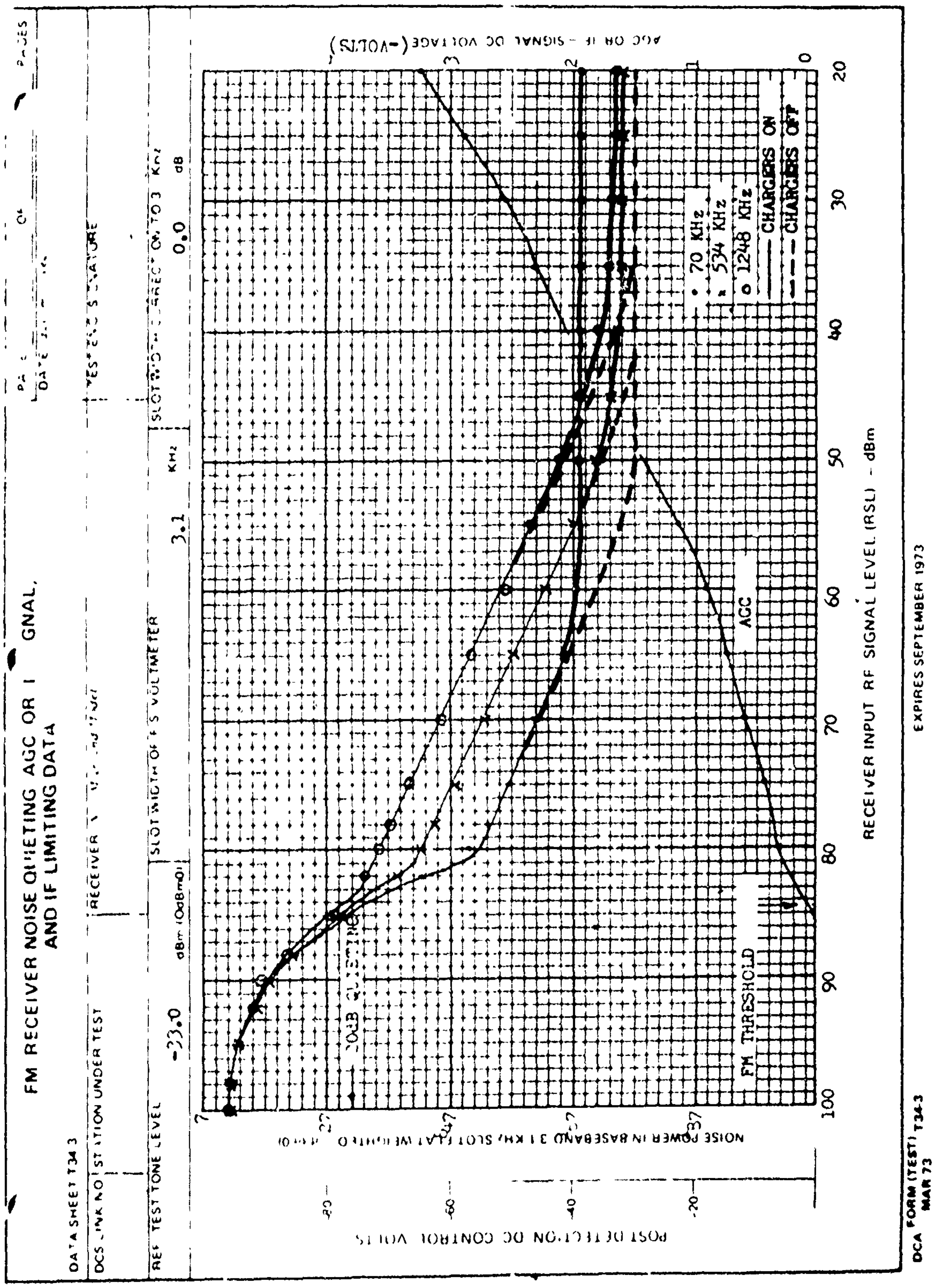

BASEBAND SUPPRESSION
LOS M/W Receiver
Note: This LOS M/W receiver is not the one used in the rest of this report.
Data for two curves taken one hour apart under identical conditions.
Horizontal scale not calibrated in absolute level.
RSL dB
-30 -20 -10 0 +10 +20 +30 +40 +50 +60 +70 +80
Signal Level dB
0 -1 -2 -3 -4 -5 -6 -7 -8 -9

Figure 46



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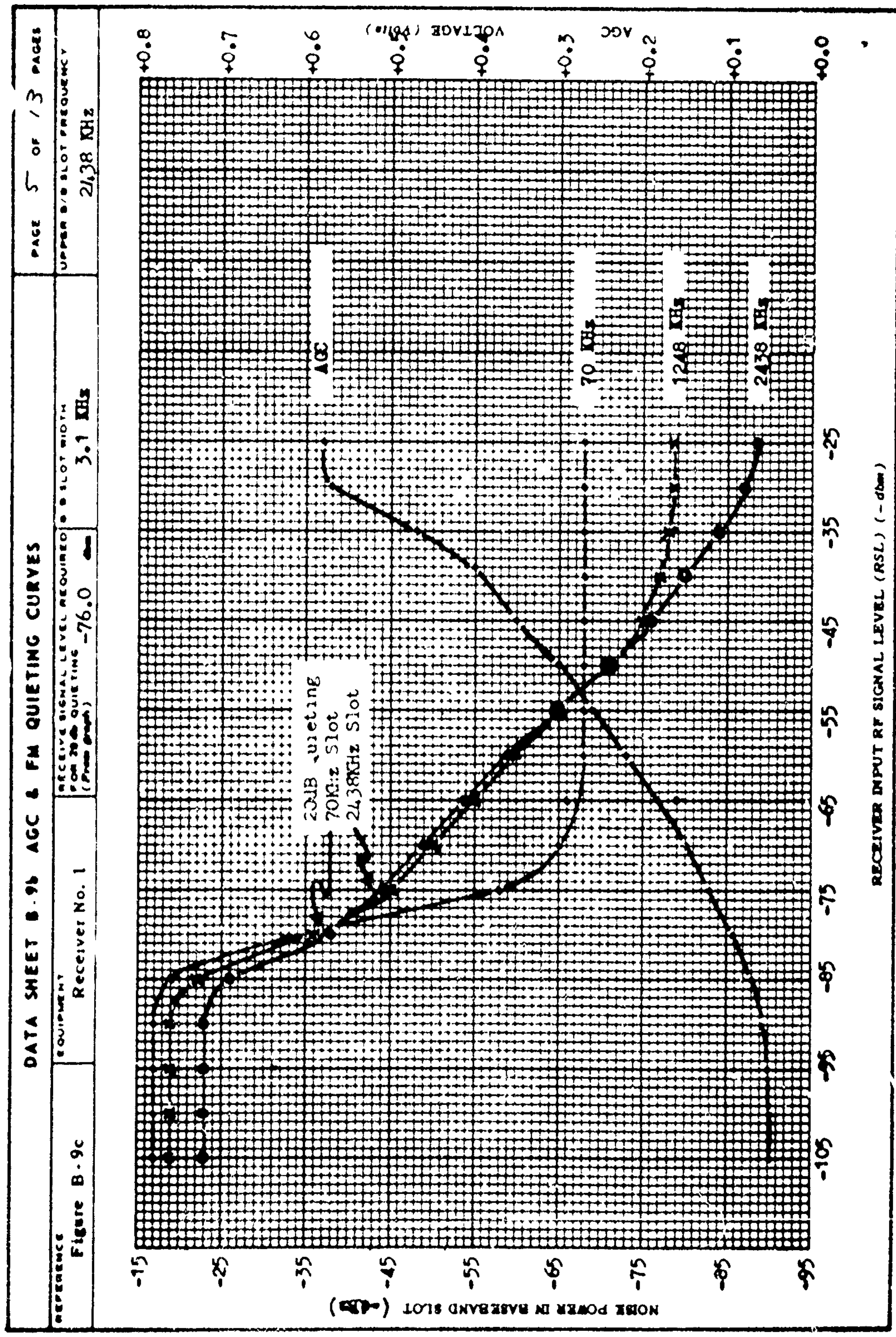
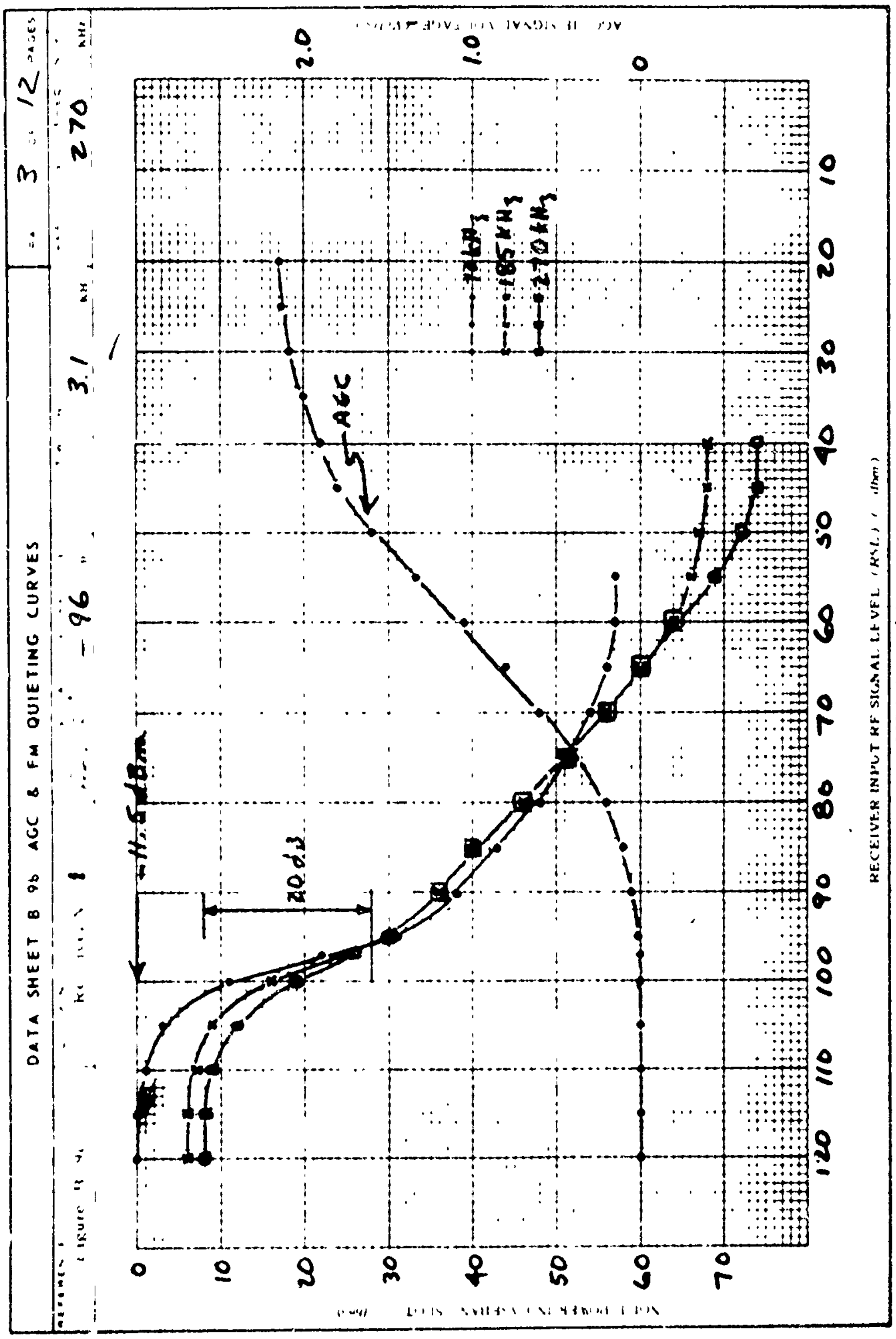
Sample Noise Quieting Curve

Figure 47



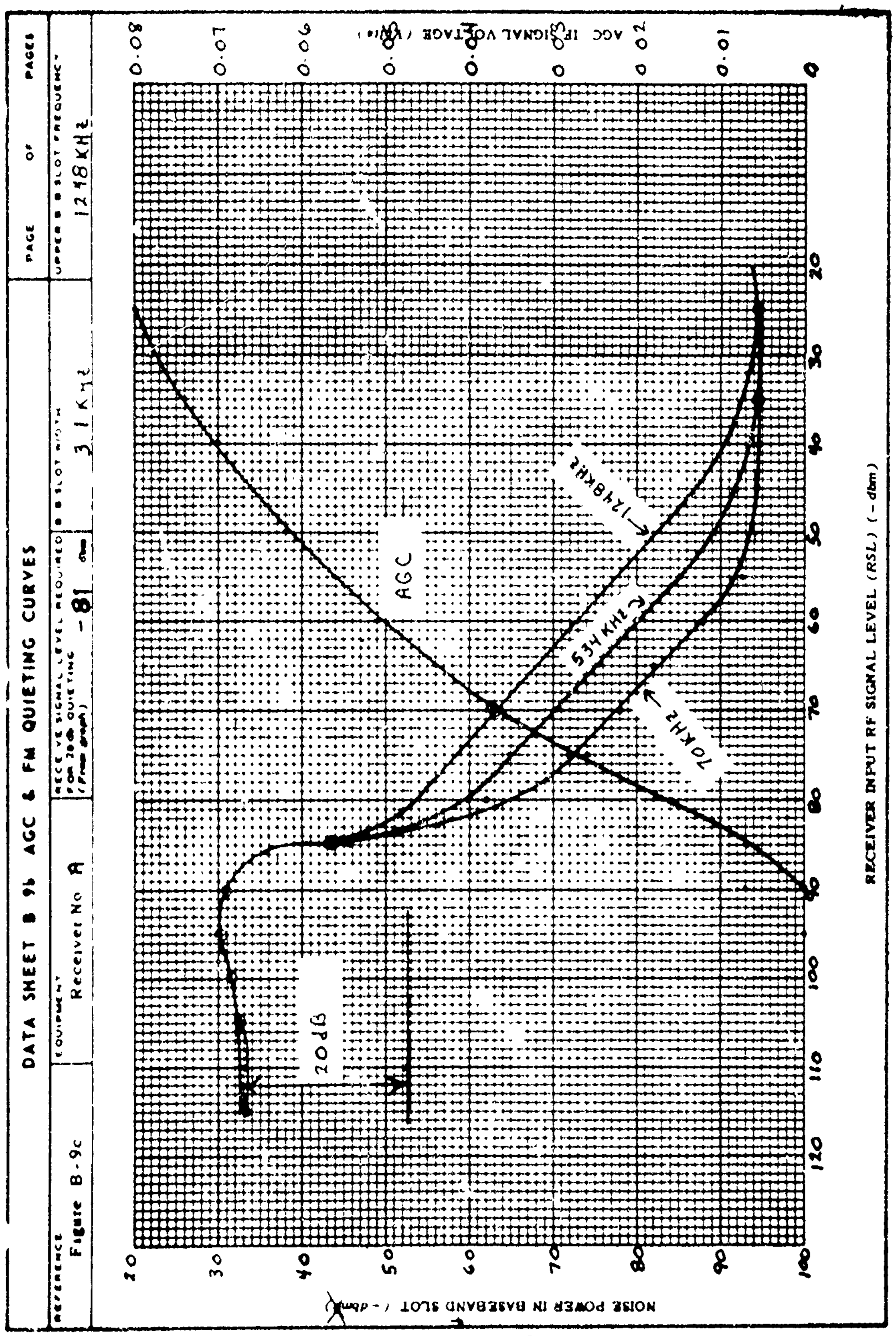
Sample Noise Quieting Curve

Figure 48



Sample Noise Quieting Curve

Figure 49



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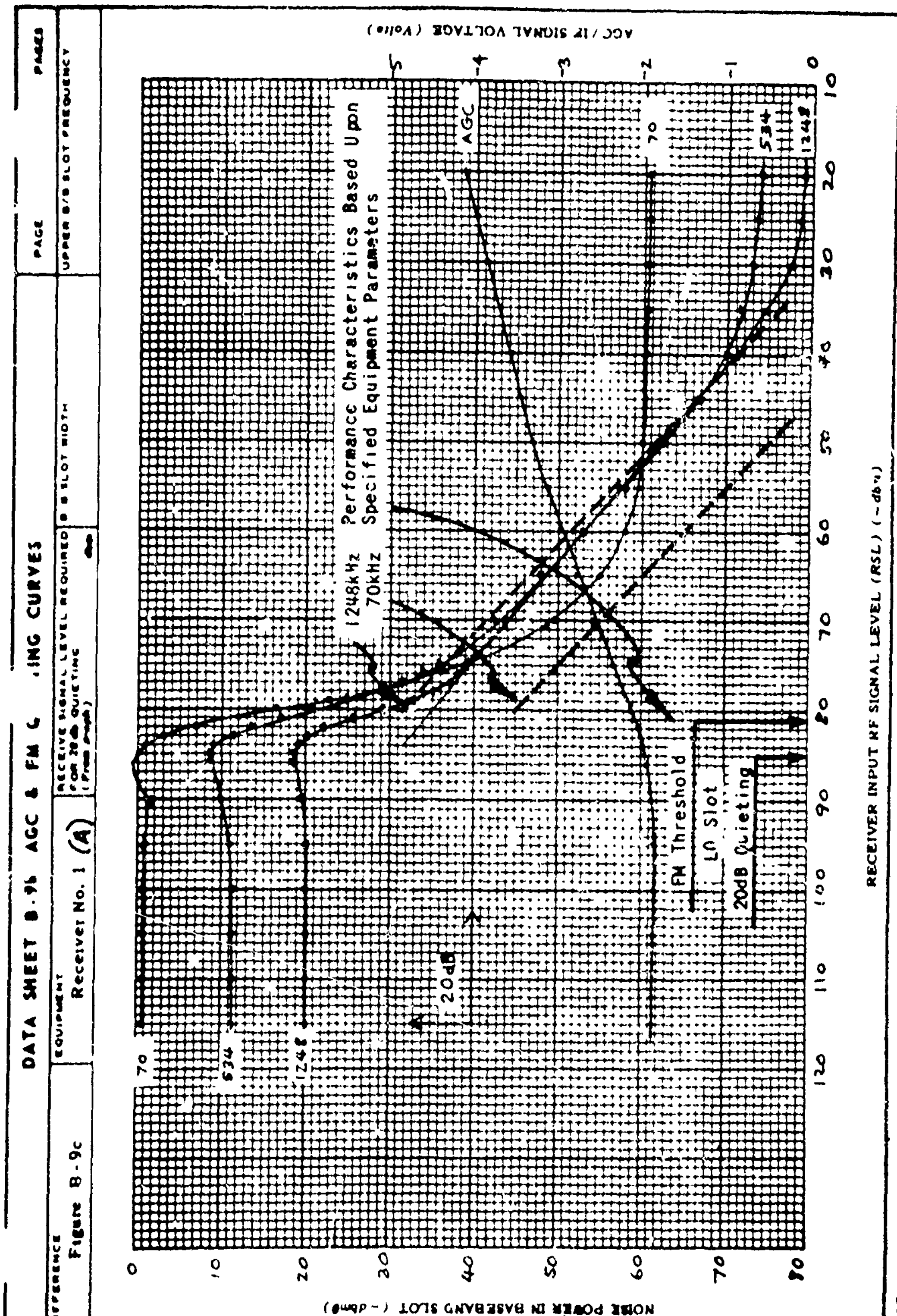
Sample Noise Quieting Curve

Figure 50



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Sample Noise Quieting Curve

Figure 51

GENERAL PURPOSE RULED DATA PAGE

DATE

Reciever 1B
PAGE OF PAGES

REFERENCE

STATION FACILITY

LINK

ORIGINATOR'S INITIALS

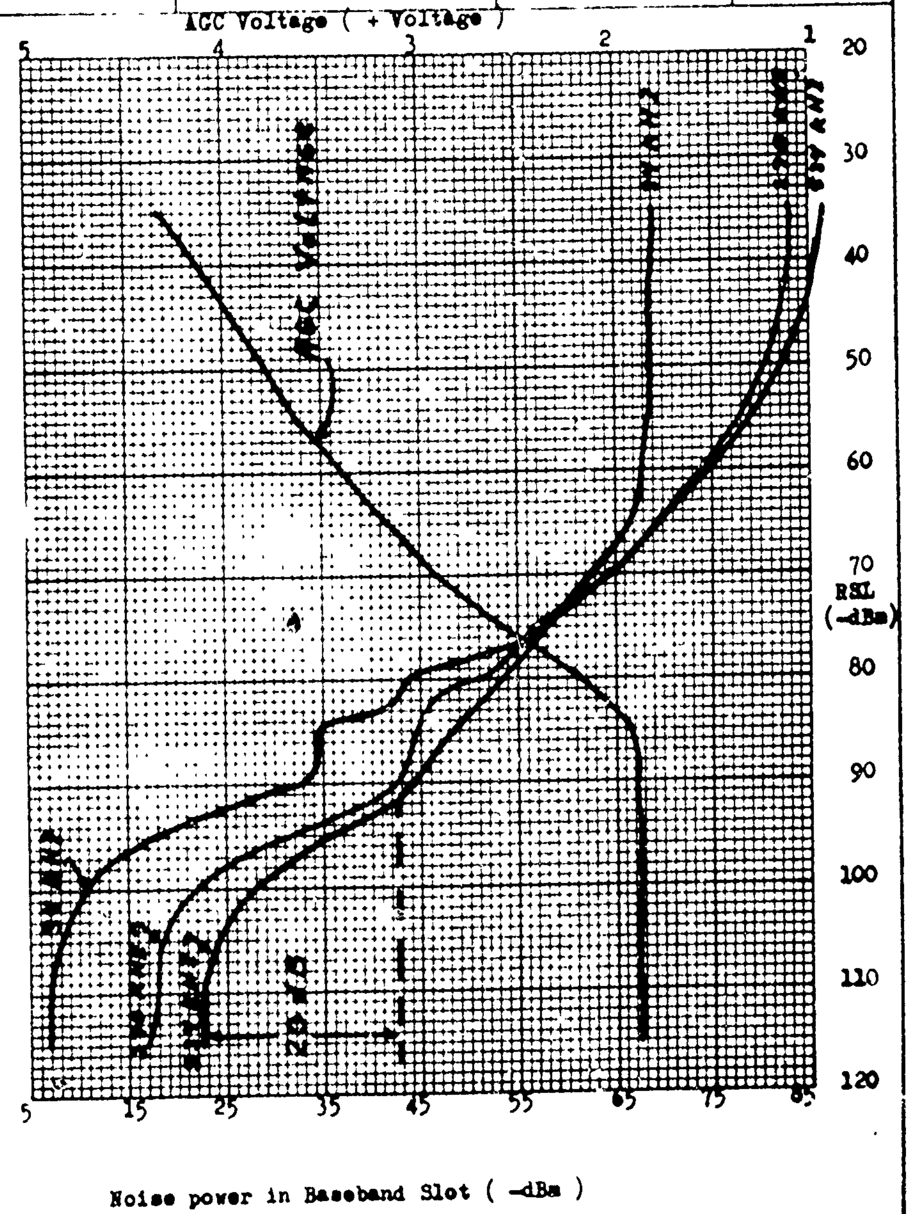


FIGURE TITLE

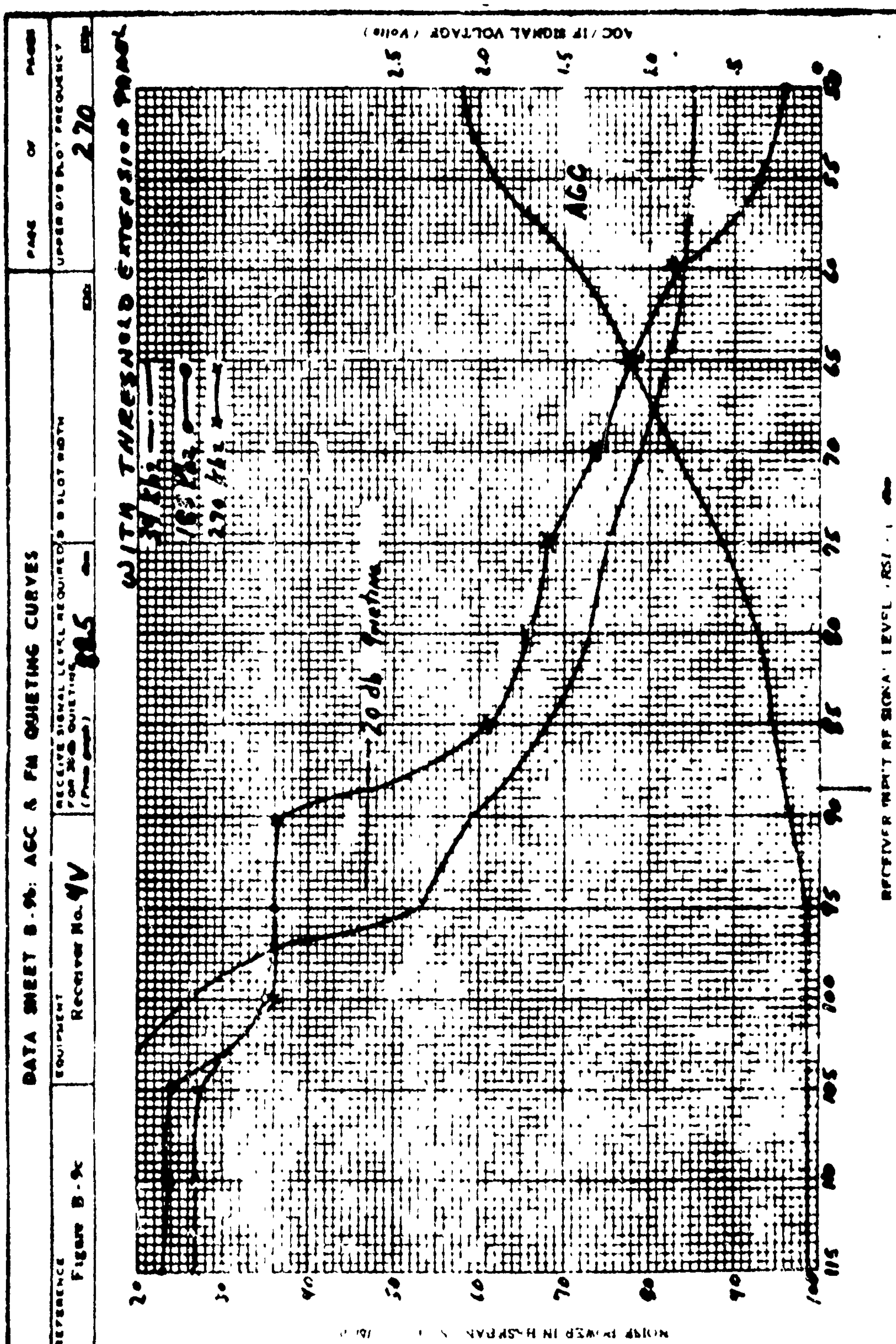
Reciever Quieting and AGC Curve

AFCS FORM 588, MAY 63

RPT PAGE NO

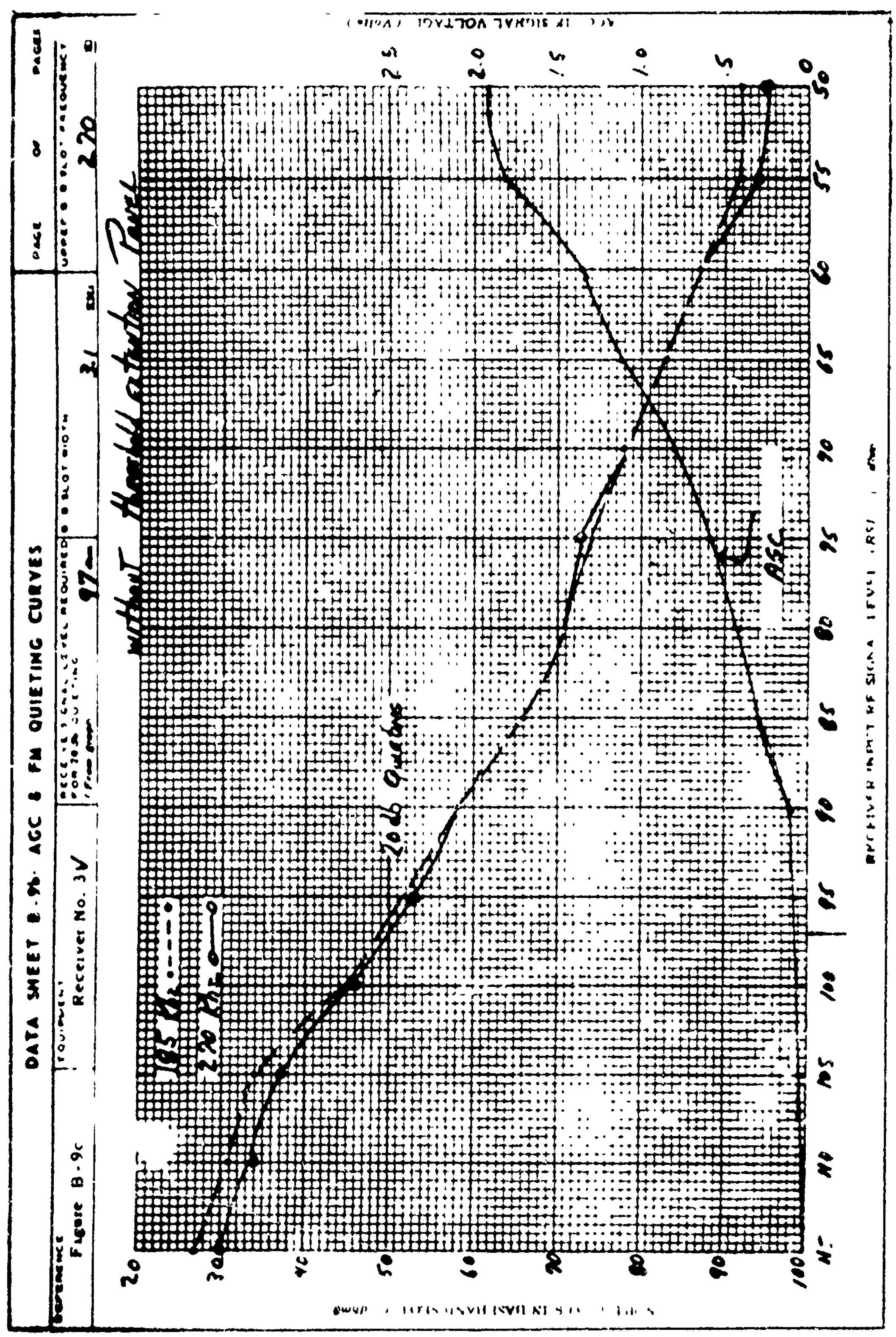
Sample Noise Quieting Curve

Figure 52



Sample Noise Quieting Curve

Figure 53



Sample Noise Quieting Curve

Figure 54

5. Predicting FM M/W Terminal Thermal Noise Performance

Symbols Used on the Following Pages:

RSL = received signal level (dBm) into receiver

NF= overall noise figure of receiver (dB) measured at the same place at which RSL is measured

B_{IF}= 3 dB bandwidth of the receiver IF

f = baseband frequency of the center of the narrow slot used (by a frequency selective voltmeter) to measure receiver baseband noise

f/B = f/B_{IF} (with f and B_{IF} converted to same units (e.g., kHz or MHz))

= baseband slot frequency compared (normalized or ratioed) to the receiver IF bandwidth

H = IF frequency response (dB) at IF center frequency plus or minus baseband slot frequency (f) minus IF frequency response (dB) at IF center frequency

= normalized IF response of the receiver which shapes the receiver thermal noise

= normalized thermal noise in a narrow slot at output of IF

h = normalized IF response of receiver (PR)

= antilog (H/10) = $10^{H/10}$

fmax = maximum baseband frequency (excluding baseband pilots)

C/N = carrier to noise ratio (dB)

=power level of received signal compared to total noise power of receiver's internally generated thermal noise which appears at the discriminator

p = carrier to noise ratio (PR)

= antilog ((C/N)/10) = $10^{(C/N)/10}$

f_p = baseband pivot frequency for the emphasis network

S = baseband test tone signal power level (dBmØ) at the receiver

S_t = baseband test tone signal power level (dBmØ) at the transmitter

N = noise power (dBmØ) in a 3.1 kHz slot located at a center frequency f in the receiver baseband

No = noise power (dBmØ) in a 3.1 KHz slot located at a very low (approximately DC) receiver baseband frequency when no carrier is applied to the input of the receiver

P = baseband pre-emphasis (dB) relative to baseband pivot frequency

Pp = baseband pre-emphasis (dB) at baseband pivot frequency relative to baseband gain (dB) at pivot frequency with pre-emphasis strapped out of the baseband circuitry

τ = emphasis time constant

$\Delta f_{/ch\ rms}$ = per channel rms deviation (kHz)

= rms deviation caused by a 0 dBmØ sine wave test tone (at baseband pivot frequency if emphasis is used)

PR = power ratio (e.g., milliwatts/milliwatts)

dB = 10 log (PR)

dBmØ = see next page

dBm = 10 log (power in milliwatts)

TLP = see next page

log (x) = common (Bragg's or base 10) logarithm of the number x

π = pi $\cong$ 3.1416

e = base of natural (Naperian) logarithms $\cong$ 2.7183

Noise Power Normalization:

5.1 In the pages which follow, all power levels will be given in artificial dBmØ values. The desirability of the dBmØ values is that they allow direct comparison of different equipment. In communication systems, power is often measured in absolute power referenced to one milliwatt. This unit is the dBm. To convert from dBmØ to dBm values and vice versa, use the following formulas:

dBmØ = dBm - TLP(dB)

dBm = dBmØ + TLP(dB)

where TLP is the Transmission Level Point in dB associated with the test point at which the measurement is made.

Receiver Baseband Signal Level:

$$S(dBm\emptyset) = S_t(dBm\emptyset) + 20 \log (1.0 - e^{-p})$$

$$C/N(dB) = RSL(dBm) + 114.0 - NF(dB) - 10 \log B_{IF}(MHz)$$

$$P(PR) = \text{antilog} ((C/N)/10) = 10^{(C/N)/10}$$

Receiver Baseband Noise in 3.1 KHz slot:

For C/N less than -10 dB:

$$N(dBm\emptyset) = No + (IF) - 4.34p$$

where the "IF" factor is obtained from the graph on the following page. The formula is listed below for certain values of f/B. The value of No is obtained using the appropriate formula listed below.

Rectangular IF

f/B = 0.0	N(dBmØ) = No - 4.34p	max. error = 0.23dB
f/B = 0.1	N(dBmØ) = No - 0.71 - 4.34p	max. error = 0.17dB
f/B = 0.2	N(dBmØ) = No - 1.43 - 4.34p	max. error = 0.06dB
f/B = 0.3	N(dBmØ) = No - 2.15 - 4.34p	max. error = 0.12dB
f/B = 0.4	N(dBmØ) = No - 2.86 - 4.34p	max. error = 0.39dB
f/B = 0.5	N(dBmØ) = No - 3.54 - 4.34p	max. error = 0.74dB

$$No(dBm\emptyset) = +29.76 - 20 \log \Delta f_{/ch\ rms}(kHz) + 10 \log B_{IF}(MHz)$$

Gaussian IF:

f/B = 0.0	N(dBmØ) = No - 4.34p	max error = 0.12dB
f/B = 0.1	N(dBmØ) = No - 0.03 - 4.34p	max. error = 0.11dB
f/B = 0.2	N(dBmØ) = No - 0.14 - 4.34p	max. error = 0.06dB
f/B = 0.3	N(dBmØ) = No - 0.31 - 4.34p	max. error = 0.00dB
f/B = 0.4	N(dBmØ) = No - 0.54 - 4.34p	max. error = 0.07dB

f/B = 0.5 N(dBmØ) = No - 0.83 - 4.34p max. error = 0.18dB

$No(dBmØ) = +29.51 - 20 \log \Delta f_{/ch\ rms}(kHz) + 10 \log B_{IF}(MHz)$

Averaged IF:

f/B = 0.0 N(dBmØ) = No - 4.34p max. error = 0.18dB

f/B = 0.1 N(dBmØ) = No - 0.37 - 4.34p max. error = 0.14dB

f/B = 0.2 N(dBmØ) = No - 0.79 - 4.34p max. error = 0.06dB

f/B = 0.3 N(dBmØ) = No - 1.23 - 4.34p max. error = 0.06dB

f/B = 0.4 N(dBmØ) = No - 1.70 - 4.34p max. error = 0.23dB

f/B = 0.5 N(dBmØ) = No - 2.19 - 4.34p max. error = 0.46dB

$No(dBmØ) = +29.64 - 20 \log \Delta f_{/ch\ rms}(kHz) + 10 \log B_{IF}(MHz)$

Note: The appropriate pre-emphasis value must be subtracted from the above values if the terminal uses emphasis.

For C/N between -10 dB and +5 dB:

$N(dBmØ) = No + (A) + (B)p + (C)p^2$

where the "A", "B", and "C" factors are obtained from the graphs on the following pages. The value of No is obtained using the appropriate formula listed below.

Rectangular IF:

f/B = 0.0 N(dBmØ) = No - 0.22 - 7.284p + 0.722p^2 max. error = 0.35 dB

f/B = 0.1 N(dBmØ) = No - 1.03 - 6.600p + 0.738p^2 max. error = 0.37dB

f/B = 0.2 N(dBmØ) = No - 1.91 - 5.036p + 0.606p^2 max. error = 0.48dB

f/B = 0.3 N(dBmØ) = No - 2.65 - 3.326p + 0.390p^2 max. error = 0.52dB

f/B = 0.4 N(dBmØ) = No - 3.26 - 1.287p + 0.011p^2 max. error = 0.49dB

f/B = 0.5 N(dBmØ) = No - 3.63 + 0.399p - 0.319p^2 max. error = 0.50dB

$No(dBmØ) = +29.76 - 20 \log \Delta f_{/ch\ rms}(kHz) + 10 \log B_{IF}(MHz)$

Gaussian IF:

$f/B = 0.0 \qquad N(dBm\emptyset) = No + 0.03 - 5.588p + 0.326p^2 \qquad max.error = 0.04dB$

$f/B = 0.1 \qquad N(dBm\emptyset) = No - 0.01 - 5.440p + 0.392p^2 \qquad max.error = 0.04dB$

$f/B = 0.2 \qquad N(dBm\emptyset) = No - 0.13 - 4.911p + 0.438p^2 \qquad max.error = 0.03dB$

$f/B = 0.3 \qquad N(dBm\emptyset) = No - 0.33 - 4.163p + 0.391p^2 \qquad max.error = 0.05dB$

$f/B = 0.4 \qquad N(dBm\emptyset) = No - 0.57 - 3.478p + 0.307p^2 \qquad max.error = 0.05dB$

$f/B = 0.5 \qquad N(dBm\emptyset) = No - 0.83 - 2.973p + 0.227p^2 \qquad max.error = 0.06dB$

$No(dBm\emptyset) = +29.51 - 20 \log \Delta f_{/ch\ rms}(kHz) + 10 \log B_{IF}(MHz)$

Averaged IF:

$f/B = 0.0 \qquad N(dBm\emptyset) = No - 0.10 - 6.426p + 0.521p^2 \qquad max.error = 0.17dB$

$f/B = 0.1 \qquad N(dBm\emptyset) = No - 0.52 - 6.014p + 0.564p^2 \qquad max.error = 0.18dB$

$f/B = 0.2 \qquad N(dBm\emptyset) = No - 1.02 - 4.969p + 0.521p^2 \qquad max.error = 0.23dB$

$f/B = 0.3 \qquad N(dBm\emptyset) = No - 1.49 - 3.743p + 0.390p^2 \qquad max.error = 0.26dB$

$f/B = 0.4 \qquad N(dBm\emptyset) = No - 1.92 - 2.379p + 0.158p^2 \qquad max.error = 0.26dB$

$f/B = 0.5 \qquad N(dBm\emptyset) = No - 2.24 - 1.282p - 0.047p^2 \qquad max.error = 0.26dB$

$No(dBm\emptyset) = +29.64 - 20 \log \Delta f_{/ch\ rms}(kHz) + 10 \log B_{IF}(MHz)$

Note: The appropriate pre-emphasis value must be subtracted from the above values if the terminal uses emphasis.

5 2 Due to the computational difficulties involved in obtaining exact results for low C/Ns, the preceeding results were obtained by fitting polynomial curves to the results of Stumpers and Rice. The maximum error in the preceeding results is graphed on the following page.

For C/N greater than + 5dB:

Rectangular IF:

$$N(dBm\emptyset) = -139.0 - 20 \log \Delta f_{/ch\ rms}(kHz) - RSL(dBm) + NF(dB) - P(dB)$$
$$+ 10 \log\left(f^2(kHz) + (0.32575)\left[\frac{\sqrt{p}\quad B_{IF}{}^2(kHz)}{e^p}\right]\right)$$

Gaussian IF:

$$N(\mathrm{dBm0}) = -139.2 - 20\log \Delta f_{/ch\ rms}(\mathrm{kHz}) - \mathrm{RSL}(\mathrm{dBm}) + \mathrm{NF}(\mathrm{dB}) - P(\mathrm{dB})$$

$$+\ 10\log\left(e^{-\pi(f/B)^2} f^2(\mathrm{kHz}) + (0.45029)\left[\frac{\sqrt{p}\ \ B_{IF}{}^2(\mathrm{kHz})}{e^p}\right]\right)$$

Averaged IF:

$$N(\mathrm{dBm0}) = -139.1 - 20\log \Delta f_{/ch\ rms}(\mathrm{kHz}) - \mathrm{RSL}(\mathrm{dBm}) + \mathrm{NF}(\mathrm{dB}) - P(\mathrm{db})$$

$$+10\log\left(h\ f^2(\mathrm{kHz}) + (0.38797)\left[\frac{\sqrt{p}\ \ B_{IF}{}^2(\mathrm{kHz})}{e^p}\right]\right)$$

for the above formulas

$$C/N(\mathrm{dB}) = \mathrm{RSL}(\mathrm{dBm}) + 114.0 - \mathrm{NF}(\mathrm{dB}) - 10\log B_{IF}(\mathrm{MHz})$$

$$p = \mathrm{antilog}\ (\ (C/N)/10)$$

Note: The above formulas do not include baseband phase/frequency noise associated with the transmitted carrier. This noise will dominate at very high C/N ratios.

The following special cases apply when the FM receiver is operating above FM (1 dB noise) threshold (C/N +13 dB) but with a low enough RSL that transmit phase/frequency noise is not significant.

Noise in 3.1 kHz noise slot:

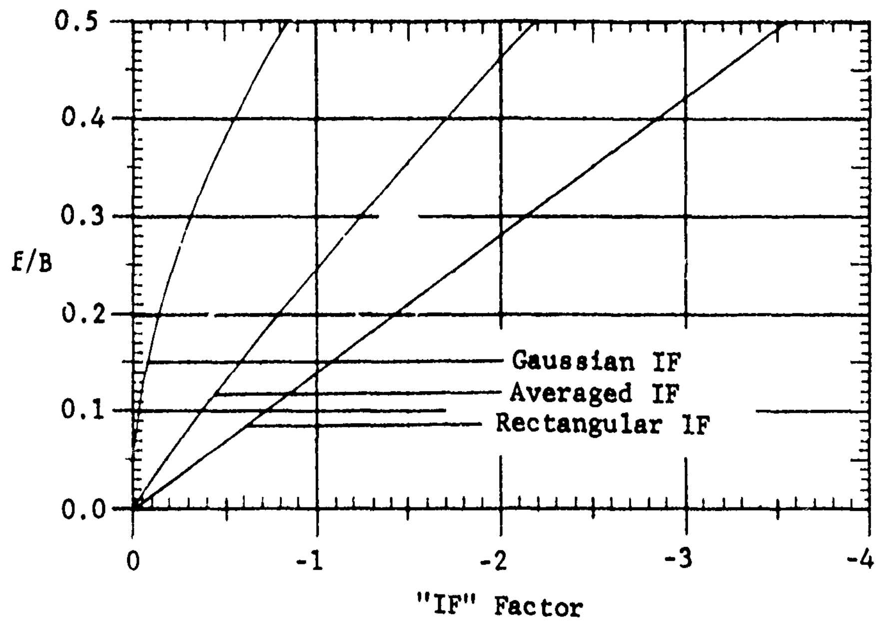
$$N(\mathrm{dBm0}) = -139.1 + 20\log f(\mathrm{kHz}) + \mathrm{NF}(\mathrm{dB}) - P(\mathrm{dB})$$
$$-20\log \Delta f_{/ch\ rms}(\mathrm{kHz}) - \mathrm{RSL}\ (\mathrm{dBm})$$

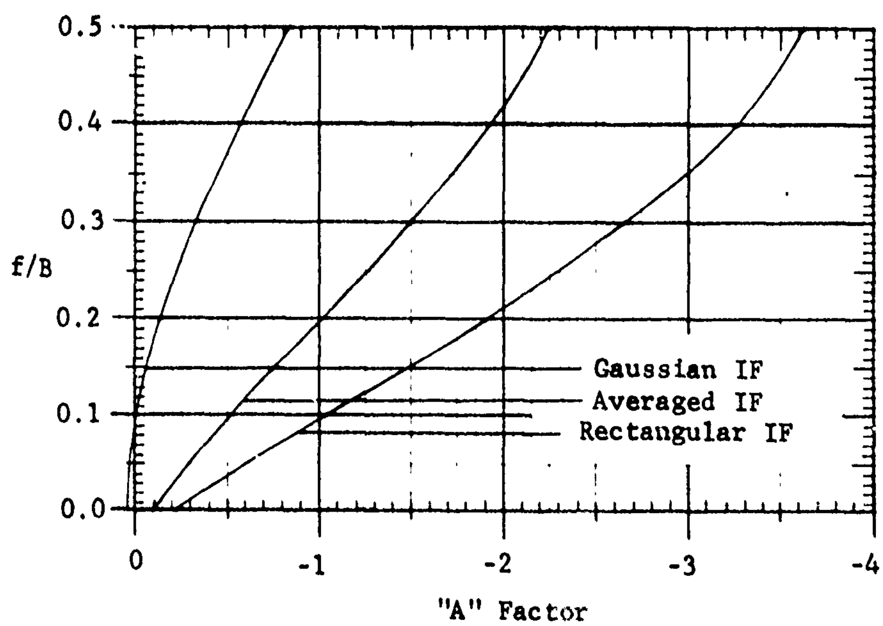
Noise in arbitrary width slot (no de-emphasis, sharp cutoff bandpass filter):

$$N(\mathrm{dBm0}) = -148.7 + 10\log\left(f^3_{max}(\mathrm{kHz}) - f^3_{min}(\mathrm{kHz})\right)$$
$$+\ \mathrm{NF}(\mathrm{dB}) - 20\log \Delta f_{/ch\ rms}(\mathrm{kHz}) - \mathrm{RSL}(\mathrm{dBm})$$

Noise in entire baseband (no de-emphasis, sharp cutoff low pass baseband filter):

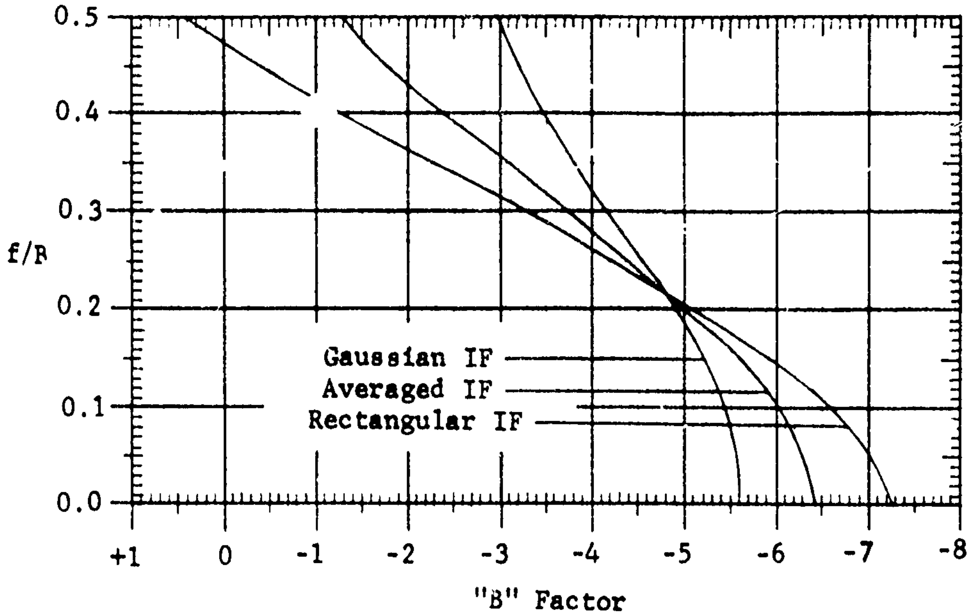
$$N(\mathrm{dBm0}) = -148.7 + 30\log f_{max}(\mathrm{kHz}) + \mathrm{NF}(\mathrm{dB})$$
$$-20\log \Delta f_{/ch\ rms}(\mathrm{kHz}) - \mathrm{RSL}(\mathrm{dBm})$$

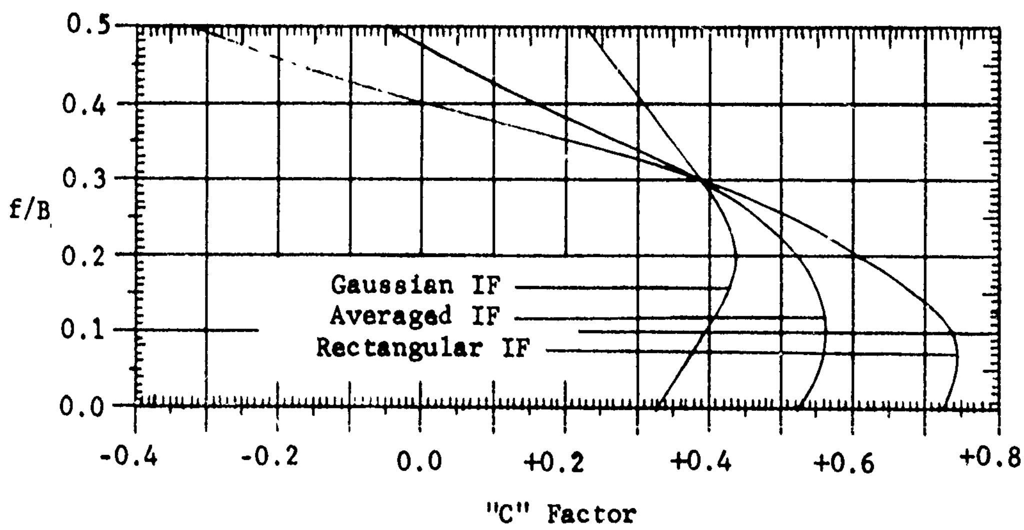




Noise Equations Factors

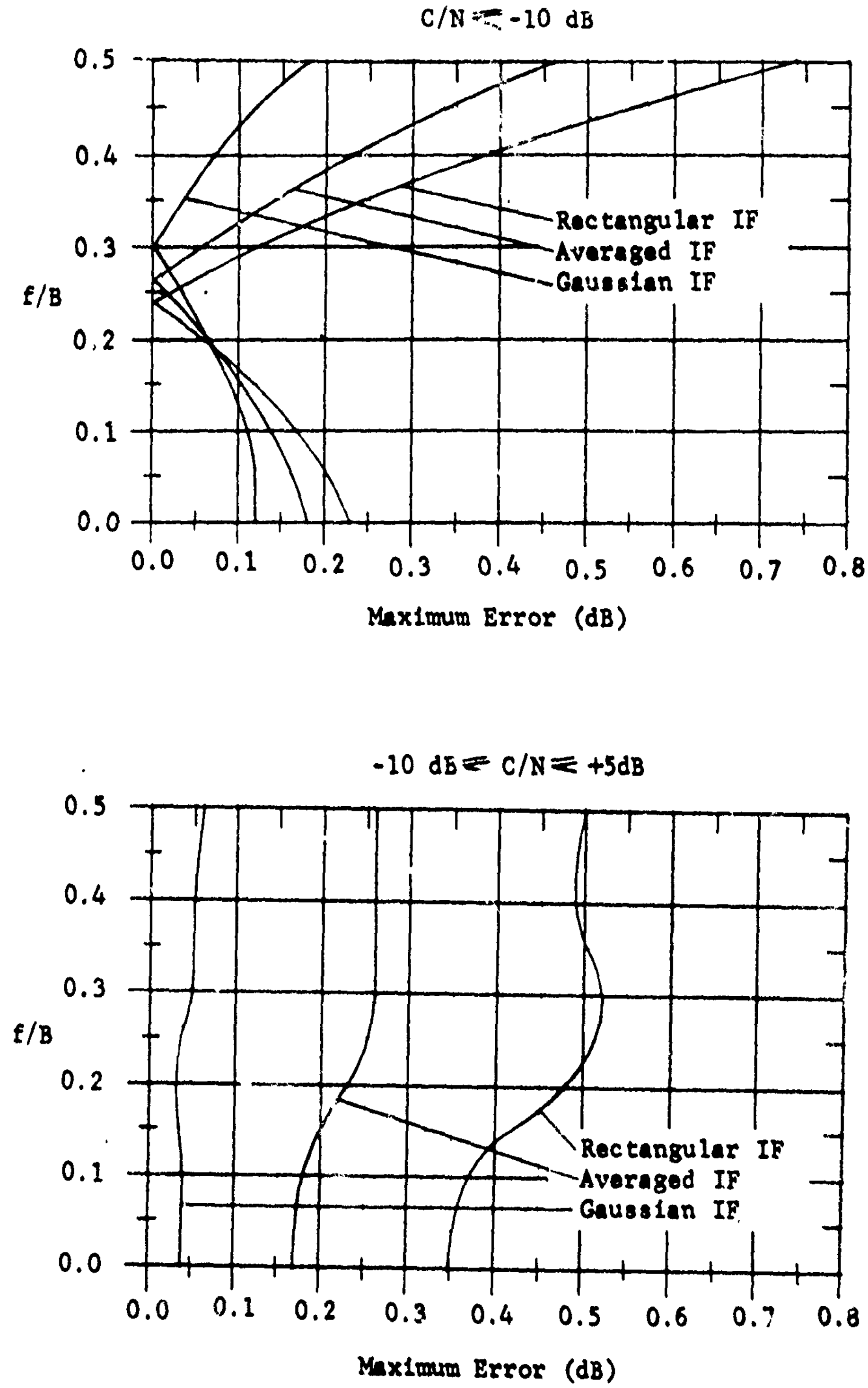
Figure 54.1





Noise Equations Factors

Figure 54.1
(cont.)



Noise Equations Error Analysis

Figure 54.2

Note that although noise in a narrow noise slot increases with the square of the slot frequency, total baseband noise increases with the cube of the upper baseband limit.

FM One dB Noise Threshold:

Narrow Noise Slots:

Rectangular IF:

$$e^{c/n}\,(f/B)^2 = (1.2580)\,(c/n)^{\frac{1}{2}}$$

Gaussian IF:

$$e^{c/n}\,(f/B)^2 = (1.7386)\,e^{+\pi(f/B)^2}\,(c/n)^{\frac{1}{2}}$$

Wide Noise Slots:

Rectangular IF:

$$e^{c/n}\left((f/B)^3\max - (f/B)^3\min\right) = (3.7741)\,(c/n)^{\frac{1}{2}}\,\left((f/B)\max - (f/B)\min\right)$$

Gaussian IF:

$$e^{c/n}\left(((f/B)^3\max - (f/B)^3\min) - (1.8849)\,((f/B)^5\max - (f/B)^5\min)\right.$$

$$+ (2.1149)\,((f/B)^7\max - (f/B)^7\min) - (1.7225)\,((f/B)^9\max - (f/B)^9\min)$$

$$\left.+ (1.1069)\,((f/B)^{11}\max - (f/B)^{11}\min)\right) = (5.2158)\,(c/n)^{\frac{1}{2}}\,\left((f/B)\max - (f/B)\min\right)$$

Where the symbols are as previously defined with

c/n = carrier to noise power ratio = antilog ((C/N)/10)
f/B = baseband slot frequency/IF bandwidth
(f/B)max = highest slot frequency/IF bandwidth
(f/B)min = lowest slot frequency/IF bandwidth

Of the last four formulas, the first three are essentially exact. The fourth is based on a truncated MacLaurin series expansion of e^x. Maximum error due to this approximation is 0.5% at f/B = 0.5. The error is considerably less for smaller f/B values.

6. Baseband Signal and Slot Noise Versus RSL Using Generalized Charts

6.1 Although quieting curves can be predicted using the preceeding formulas, the process is cumbersome. To facilitate quieting curve prediction normalized charts have been produced based on various FM receiver IF characteristics. To predict the FM noise quieting curve for an FM receiver, a normalized quieting curve can be chosen based **on** the FM receiver's IF characteristic. If the FM receiver's IF response is unknown, the averaged IF response curve is suggested as a choice. The normalized dB values for slot noise and carrier to noise ratio (C/N) can be converted to dBmØ and dBm values respectively using the following procedures.

6.2 The slot noise measured by the frequency selective voltmeter (FSV) is plotted on the vertical scale. The noise corresponding to 0 dB on the slot noise scale is determined from the following formulas:

$$N(\text{dBmØ}) = +29.6 - 20 \log \Delta f_{/ch\ rms}(\text{kHz}) + 10 \log B_{IF}(\text{MHz}) + CF(\text{dB})$$

$$CF(\text{dB}) = 10 \log \left(\frac{\text{3dB bandwidth of FSV(kHz)}}{\text{3.1 kHz}}\right)$$

= correction factor to allow for actual noise measurement bandwidth (tabulated on next page).

6.3 Use of the correction factor CF produces the dBmØ noise values which will be read by the FSV.

6.4 The actual RSL is plotted on the horizontal scale. The RSL corresponding to 0 dB on the C/N scale is found from the following formulas:

$$RSL(\text{dBm}) = -114.0 + NF(\text{dB}) + 10 \log B_{IF}(\text{MHz})$$

6.5 The slot noise is obtained for a given RSL by going up vertically from the RSL to the appropriate (f/B) line. From this line, go to the left horizontally and read noise from dBmØ scale.

6.6 The appropriate (f/B) valve is found by dividing the baseband slot frequency (frequency to which FSV is tuned) f (in kHz) by the receiver IF bandwidth (in kHz).

$$(f/B) = \frac{\text{Baseband slot frequency (kHz)}}{\text{IF bandwidth (kHz)}}$$

6.7 If the desired (f/B) valve is between the (f/B) valves listed on the curves, read the slot noise N for the (f/B) value which is closer to the desired (f/B) value and add the following correction factor to the noise valve read from the graph. Theoretical FM noise threshold has been indicated on the graphs by short vertical lines.

Theoretical FM Noise Quieting Curves

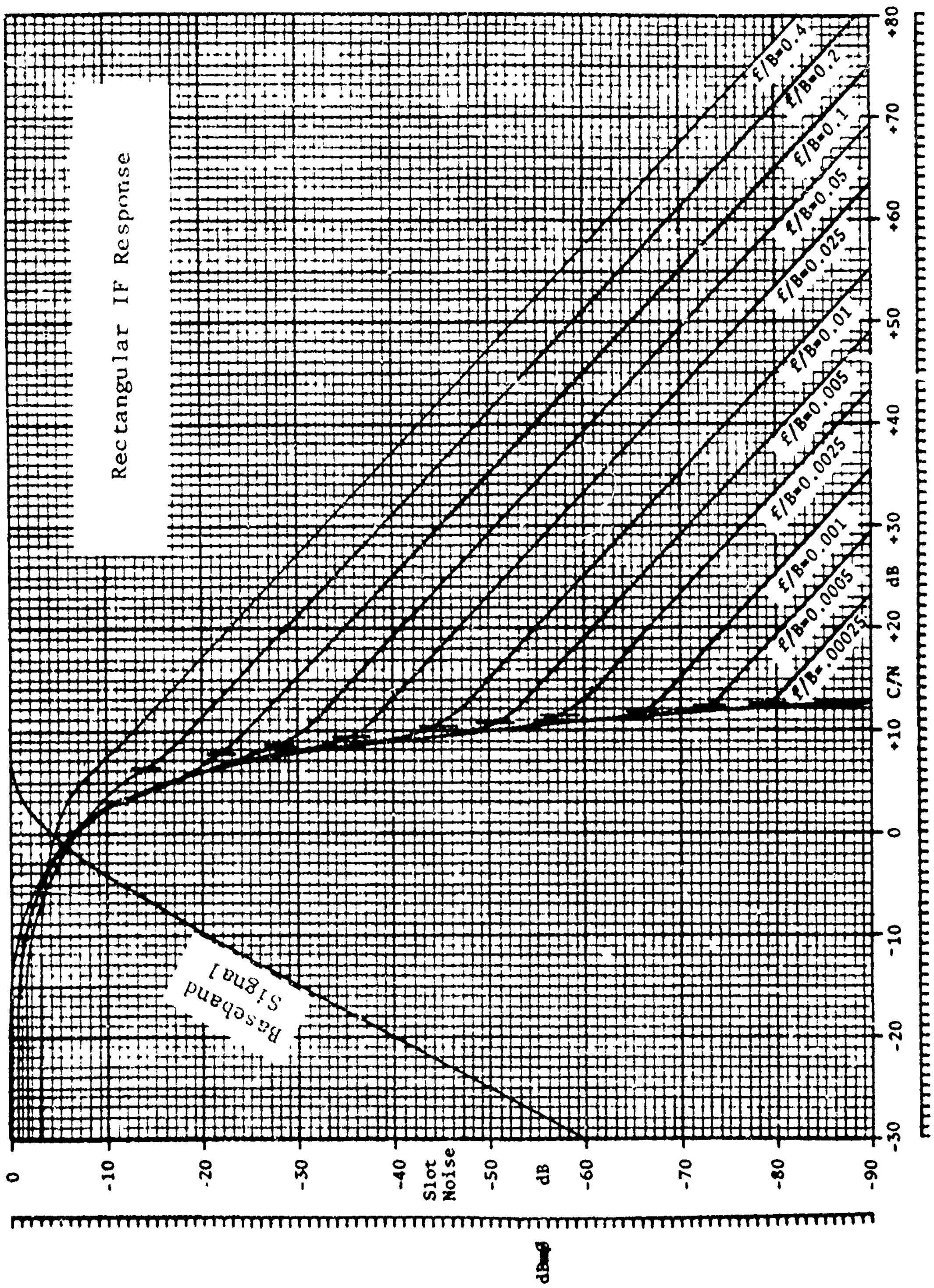


Figure 55

Theoretical FM Noise Quieting Curves

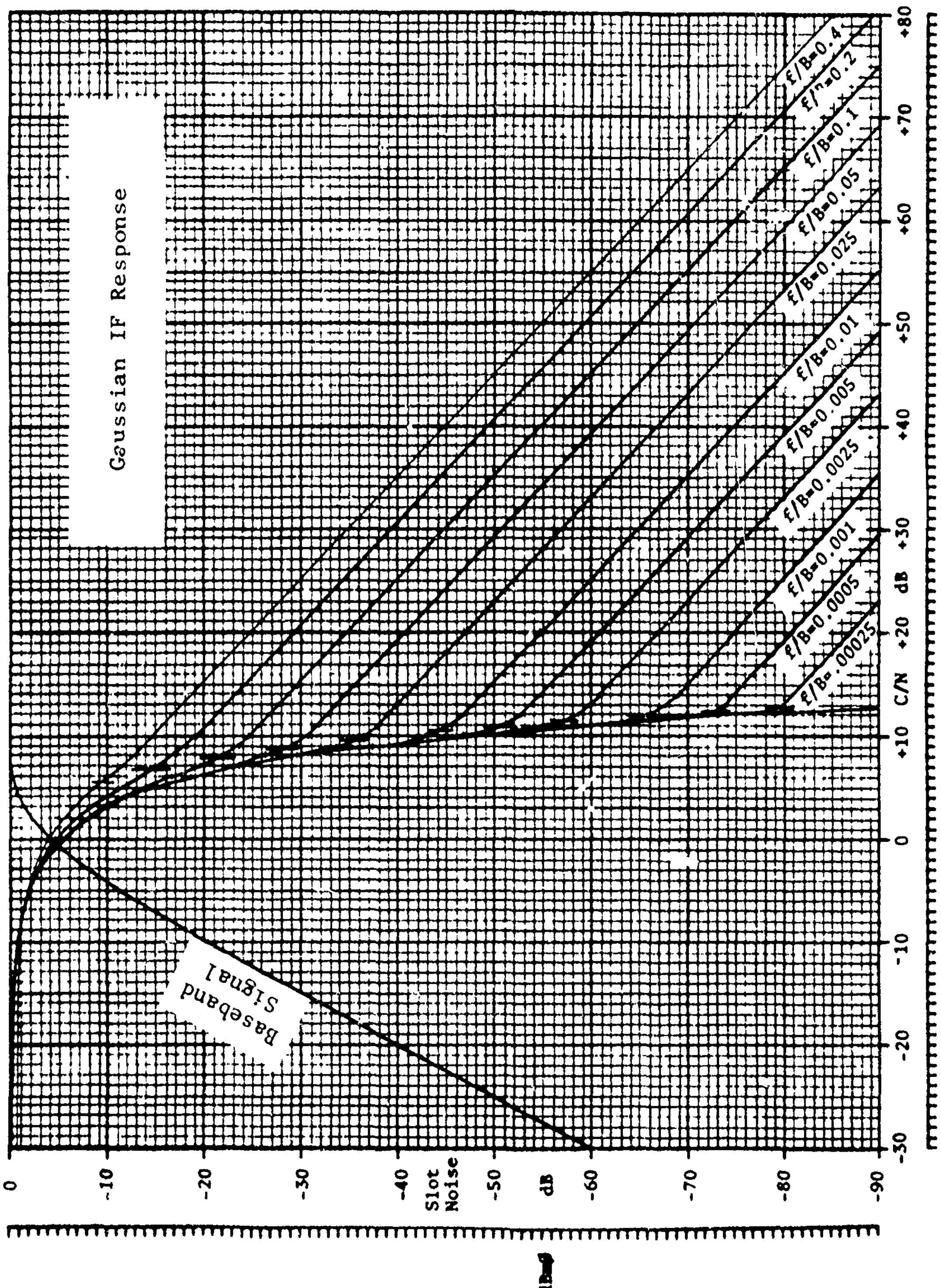


Figure 56

Theoretical FM Noise Quieting Curves

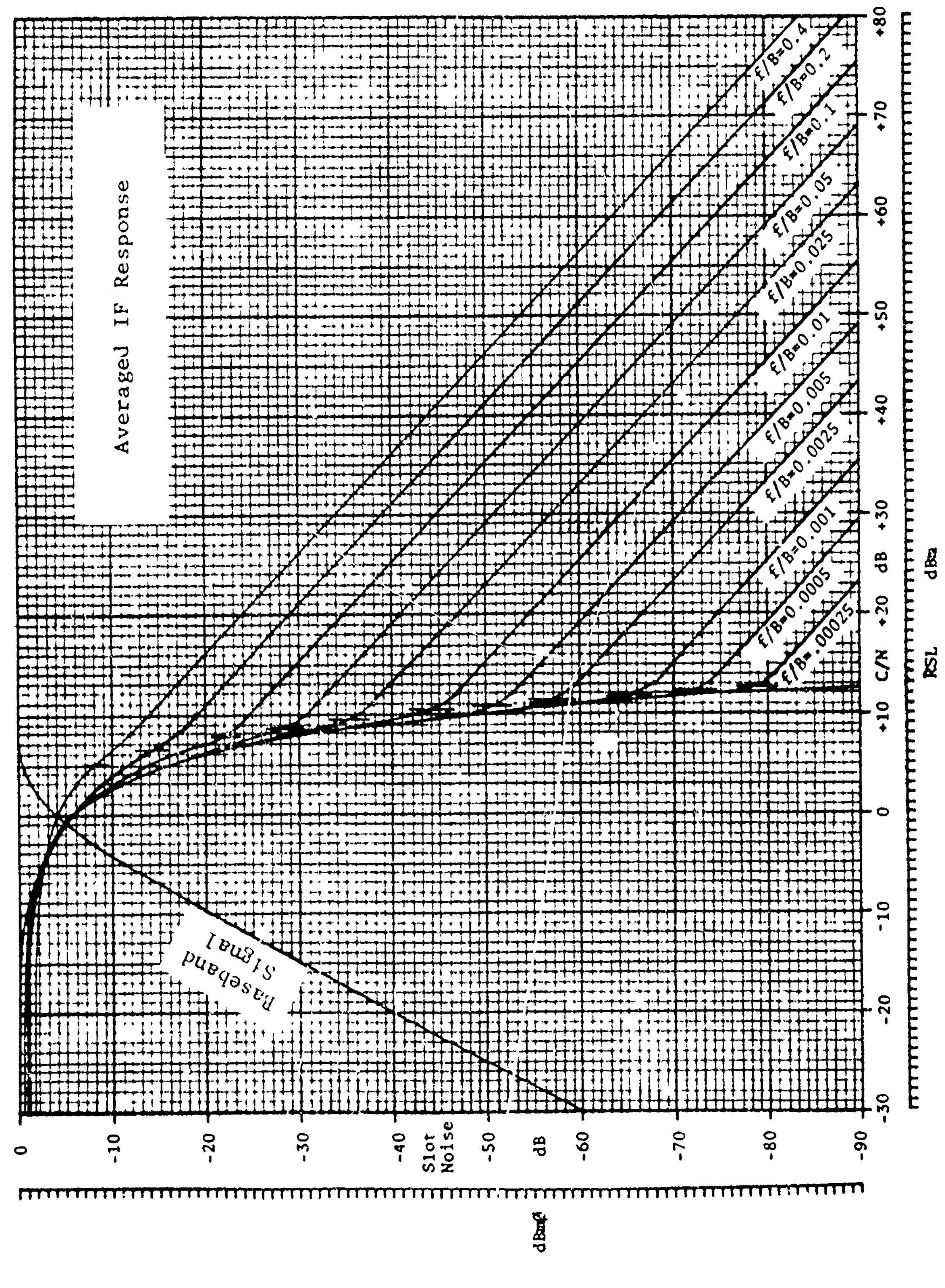


Figure 57

Measurement Bandwidth (Hz)	Conversion Factor (dB)	Measurement Bandwidth (Hz)	Conversion Factor (dB)
50K	-12.08	2.0K	+ 1.90
48K	-11.90	1.74K	+ 2.51
20K	- 8.10	1.0K	+ 4.91
15K	- 6.85	500	+ 7.92
10K	- 5.09	300	+10.14
5K	- 2.08	200	+11.90
4K	- 1.11	100	+14.91
3.1K	0.00	50	+17.92
3.0K	+ 0.14	20	+21.90
2.3K	+ 1.30	10	+24.91

Measurement Bandwidth to 3.1kHz Bandwidth Conversion Factors

(Uniformly distributed noise power assumed)

Table 19

$$N_{cf} = 20 \log \left(\frac{(f/B)\ \text{desired}}{(f/B)\ \text{from graph}} \right)$$

6.8 To evaluate wideband FM receiver performance, it is desirable to measure receiver noise quieting in at least three noise slots. To facilitate later comparison with Basic Intrinsic Noise Ratio (BINR) and Noise Power Ratio (NPR) measurements taken with the same receiver, it is recommended that quieting curves be taken at the same slot frequencies as those used for BINR/NPR measurements. Recommended frequencies are listed on the following page. The listed frequencies were chosen to conform to current CCIR recommendations and the recommendations of Tent. If there **were a conflict, the CCIR values were chosen.**

6.9 As an example, consider finding slot noise for an RSL corresponding to a +35 dB C/N. Refer to the following graph. Assume the (f/B) desired value is (0.0706).

$$N(f/B = 0.1) = -50.0\text{dB}$$

$$N(f/B = 0.0706) = N(f/B = 0.1) + 20 \log \frac{0.0706}{0.1}$$

$$= -50.0 + (-3.0) = -53.0\text{dB}$$

Alternately

$$N(f/B = 0.05) = -56.0\text{dB}$$

$$N(f/B = 0.0706) = N(f/B = 0.05) + 20 \log \frac{0.0706}{0.05}$$

$$= -56.0 + (3.0) = -53.0\ \text{dB}$$

6.10 The baseband signal level is a function of RSL near FM threshold. The signal level is found by reading the correction factor S_{CF} from the slot noise (dB) scale on the graph on the next page. The level (disregarding noise) of the received baseband signal S is found by using the following formula:

$$S(\text{dBm}Ø) = S_t(\text{dBm}Ø) + S_{CF}(\text{dB})$$

6.11 For example, at an RSL corresponding to OdB C/N, the signal level will have dropped 4 dB ($S = S_t - 4$ dB).

6.12 If the receiver being tested has de-emphasis, the slot noise power read from the graphs must be modified by the effect of the de-emphasis networks. To determine the actual noise measured by the FSV after the noise has passed through the de-emphasis networks, subtract (algebraicly)

Number of Channels	Baseband Frequencies (kHz)	Slot Frequencies (kHz)		
12	12-60	27	40	56
24	12-108	40	70	105
36	12-156	40	70	105
48	12-204	40	105	185
	*60-252	70	185	245*
60	12-252	40	185	245
	60-300	70	185	270
72	*12-300	40	185	270*
120	60-552	70	270	534
132	*12-552	40	270	534*
240	60-1052	70	534	1002
300	60-1300	70	534	1248
420	*60-1796	70	534	1248*
600	60-2660	70	1248	2438
900	316-4188	534	2438	3886
960	60-4028	70	2438	3886
1200	316-5564	534	3886	5340
1260	*60-5564	70	3886	5340*
	*60-5636	70	3886	5340*
1800	316-8204	534	3886	7600
2400	*316-11404	534	3886	7600*
2700	316-12388	534	3886	11700

Recommendations conform to CCIR NPR/BINR slot frequencies.
* No CCIR recommendations available for these basebands. *

Recommended Quieting Curve Slot Frequencies

Table 20

Theoretical FM Noise Quieting Curves

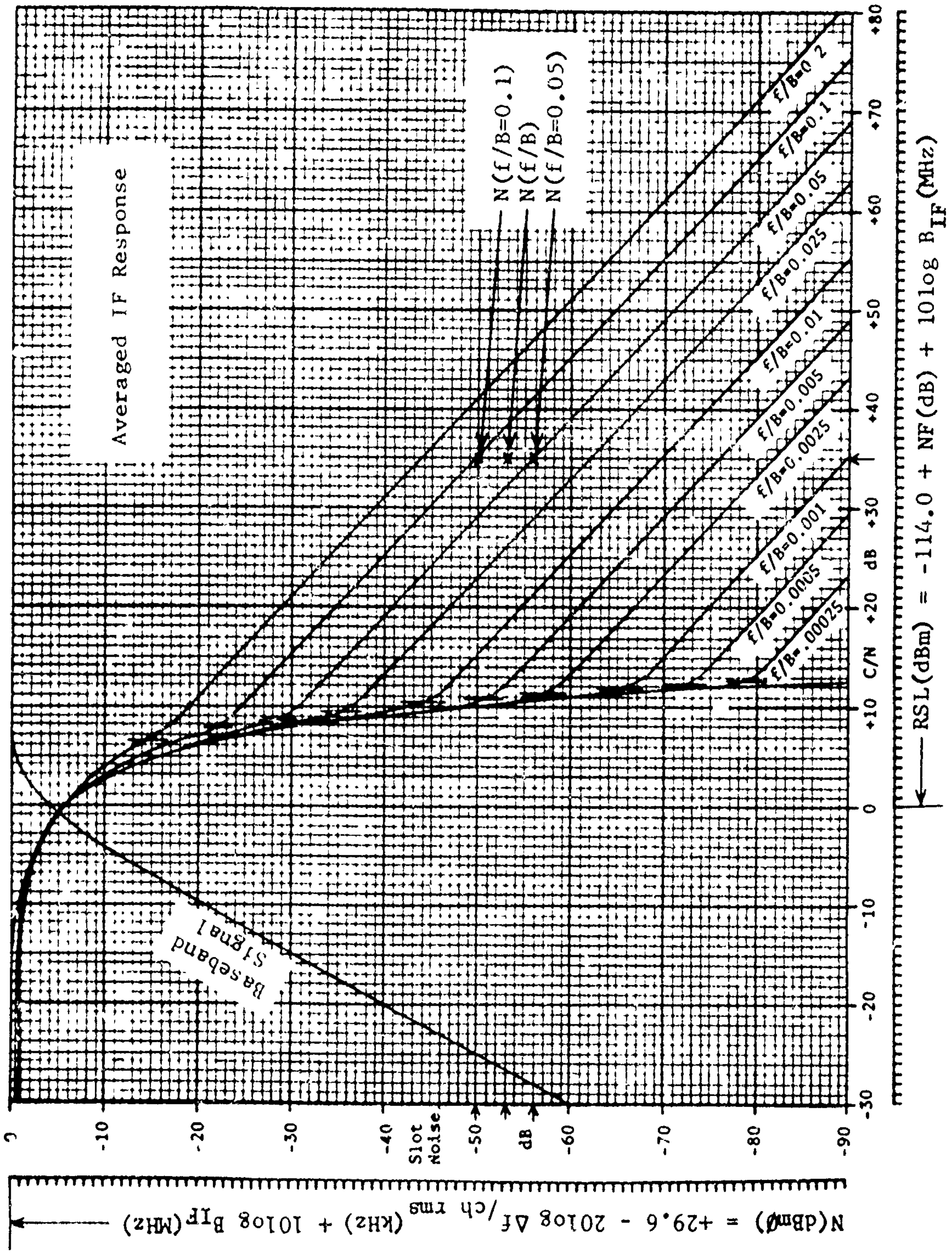


Figure 58

Theoretical Baseband Signal Suppression

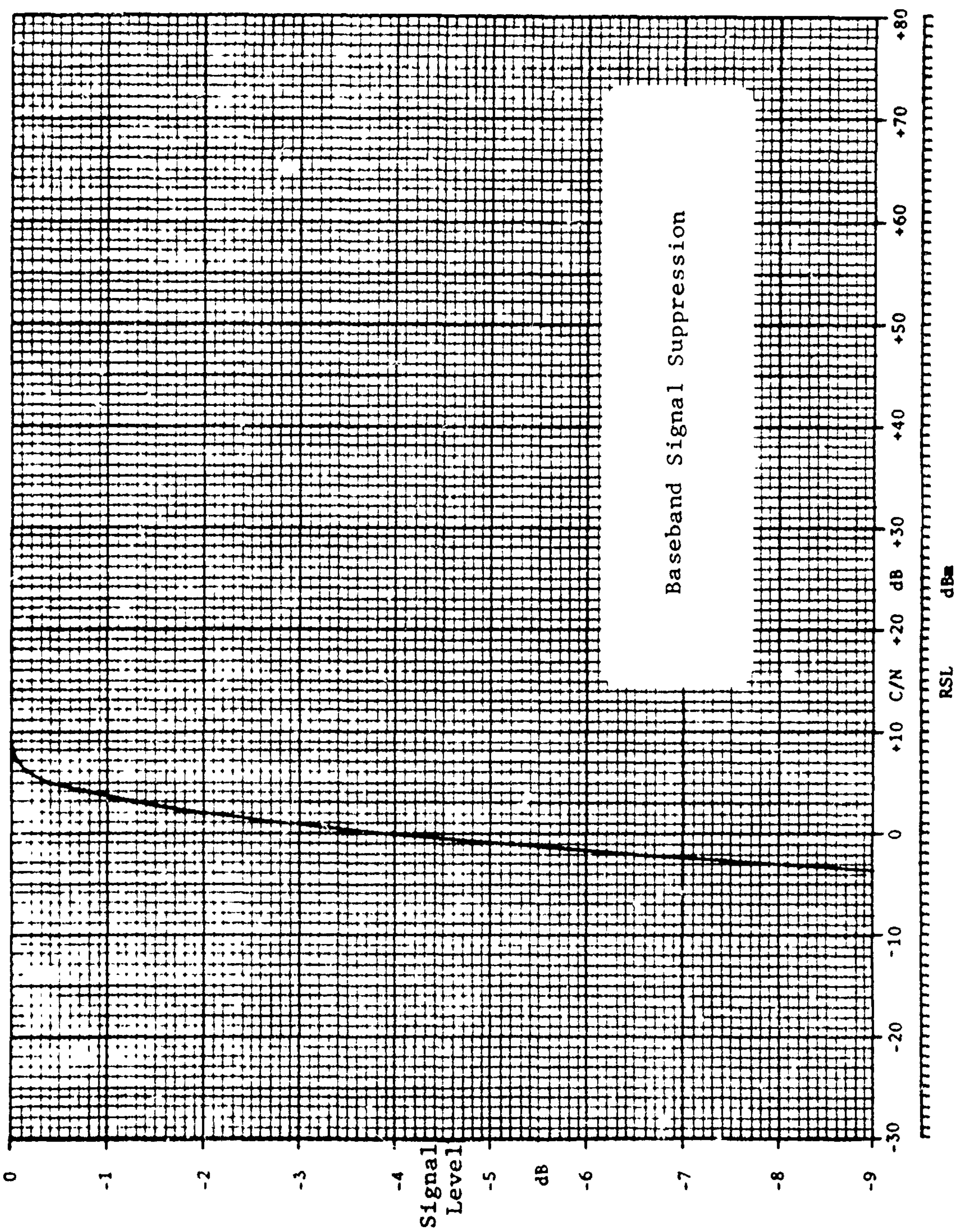


Figure 59

the pre-emphasis effect (dB) for the slot frequency of interest $\bar{f}$ relative to the baseband pivot frequency. The formulas for the pre-emphasis effect valves for the various types of pre-emphasis networks are listed on the following pages. For CCIR and EIA emphasis networks, the pre-emphasis valve may be read directly from the following table.

General Pre-emphasis Networks:

CCIR:

$$P(dB) = 10 \log (+0.400 + 1.20\,(f/fmax)^2 + 0.801\,(f/fmax)^4 + 1.03\,(f/fmax)^6 - 0.913\,(f/fmax)^8)$$

$Pp(dB) = 0.0$ dB

$fp = (0.608)\, fmax$

EIA:

$$P(dB) = 10 \log (0.250 + 2.25\,(f/fmax)^2)$$

$Pp(dB) = 0.0$ dB

$fp = (0.577)\, fmax$

Time Constant:

$$P(dB) = 10 \log (1.0 + 39.48\,\tau^2 f^2) - 10 \log (1.0 + 13.16\,\tau^2 fmax^2)$$

$$Pp(dB) = 10 \log (1.0 + 13.16\,\tau^2 fmax^2)$$

$fp = (0.577)\, fmax$

τ = time constant (μsec)
f = baseband frequency (kHz)
fmax = maximum baseband frequency (kHz)

REL Pre-emphasis Networks:

72 channel system:
(12 to 300 KHz baseband)

$$P(dB) = 10 \log (0.1743 + 2.477\,(f/fmax)^2)$$

$Pp(dB) = +7.58$

fp = (0.577) fmax = 173 kHz

132 channel system:
(12 to 552 kHz baseband)

$$P(dB) = 10 \log (0.1456 + 2.563 (f/fmax)^2)$$

Pp(dB) = +8.37

fp = (0.577 fmax) = 319 kHz

252 channel system
(12 to 1052 KHz baseband)

$$P(dB) = 10 \log (0.160 + 2.520 (f/fmax)^2)$$

Pp(dB) = +7.95

fp = (0.577) fmax = 607 kHz

Limitations

6.13 The previous sections have indicated the slot noise produced at the baseband of an FM receiver due to the FM demodulation noise process. The implicit assumption was that the demodulator output was only proportional, regardless of C/N, to the time derivative of the phase of the composite received signal plus noise waveform. It was also assumed that the total noise power N (producing the C/N) is due to the thermal noise passing through the IF amplifier and will be applied, without further bandlimiting, to the FM demodulator. These assumptions are quite reasonable for conventional FM demodulator with hard limiting. However, the assumptions do not hold for FM demodulators with feedback (FMFB) and phase-locked loop (PLL) demodulators.

6.14 Receivers employing FMFB or PLL FM demodulators have exactly the same noise characteristics in regions C and D as any comparable conventional FM receiver. However, the FMFB and PLL receivers have improved FM noise threshold performance. The C/N at which FM threshold occurs can not be readily generalized. However, Schwartz suggests a simplification for wideband FMFB receivers which implies a maximum attainable FM threshold improvement of 7 dB for a receiver with maximum baseband frequency to IF bandwidth ratio of 0.1 (f/B = 0.1) relative to conventional FM demodulation.

6.15 The noise performance of an FMFB or PLL receiver in regions A and B is difficult to analyze. It involves not only analysis of the FM noise process, but also analysis of the loop dynamics of the FMFB or PLL demodulators under nonlinear conditions. This problem is formidable. Enloe mentions that as of 1962 the problem had not been solved.

f/f_{max}	CCIR (dB)	EIA (dB)
0.	-3.98	-6.00
0.025	-3.98	-5.98
0.050	-3.95	-5.91
0.075	-3.91	-5.79
0.100	-3.86	-5.63
0.125	-3.78	-5.43
0.150	-3.70	-5.20
0.175	-3.60	-4.95
0.200	-3.48	-4.67
0.225	-3.35	-4.37
0.250	-3.21	-4.06
0.275	-3.05	-3.75
0.300	-2.89	-3.43
0.325	-2.71	-3.10
0.350	-2.52	-2.77
0.375	-2.32	-2.45
0.400	-2.11	-2.13
0.425	-1.89	-1.81
0.450	-1.66	-1.50
0.475	-1.43	-1.19
0.500	-1.19	-0.88
0.525	-0.93	-0.59
0.550	-0.68	-0.29
0.575	-0.42	-0.01
0.600	-0.15	0.27
0.625	0.12	0.55
0.650	0.40	0.81
0.675	0.68	1.07
0.700	0.96	1.33
0.725	1.24	1.58
0.750	1.52	1.82
0.775	1.79	2.06
0.800	2.07	2.30
0.825	2.34	2.53
0.850	2.61	2.75
0.875	2.86	2.97
0.900	3.12	3.18
0.925	3.36	3.39
0.950	3.59	3.60
0.975	3.80	3.80
1.000	4.01	4.00

CCIR/EIA Pre-emphasis Values
(change dB sign for de-emphasis value)

Table 21

f/f_{max}	72 Ch. (dB)	132 Ch. (dB)	252 Ch. (dB)
0.	-7.59	-8.37	-7.96
0.025	-7.55	-8.32	-7.92
0.050	-7.44	-8.18	-7.79
0.075	-7.25	-7.96	-7.59
0.100	-7.01	-7.66	-7.32
0.125	-6.72	-7.31	-7.00
0.150	-6.38	-6.92	-6.64
0.175	-6.02	-6.50	-6.25
0.200	-5.63	-6.05	-5.84
0.225	-5.23	-5.60	-5.41
0.250	-4.83	-5.15	-4.98
0.275	-4.42	-4.69	-4.55
0.300	-4.01	-4.25	-4.13
0.325	-3.61	-3.81	-3.70
0.350	-3.21	-3.38	-3.29
0.375	-2.82	-2.96	-2.89
0.400	-2.44	-2.55	-2.49
0.425	-2.06	-2.16	-2.11
0.450	-1.70	-1.77	-1.74
0.475	-1.35	-1.40	-1.38
0.500	-1.00	-1.04	-1.02
0.525	-0.67	-0.70	-0.68
0.550	-0.35	-0.36	-0.35
0.575	-0.03	-0.03	-0.03
0.600	0.28	0.29	0.28
0.625	0.58	0.59	0.59
0.650	0.87	0.89	0.88
0.675	1.15	1.18	1.17
0.700	1.42	1.47	1.45
0.725	1.69	1.74	1.72
0.750	1.95	2.01	1.98
0.775	2.21	2.27	2.24
0.800	2.45	2.52	2.49
0.825	2.70	2.76	2.73
0.850	2.93	3.00	2.97
0.875	3.16	3.24	3.20
0.900	3.39	3.47	3.43
0.925	3.61	3.69	3.65
0.950	3.82	3.91	3.86
0.975	4.03	4.12	4.07
1.000	4.23	4.33	4.28

REL Pre-emphasis Values
(change dB sign for de-emphasis value)

Table 21.1

6.16 In general, the analysis developed in the previous sections of this report will describe the noise performance of FMFB and PLL demodulators in regions C and D. The theoretical FM threshold C/N values will be pessimistic and the theoretical slot noise for regions A and B do not apply for FMFB or PLL demodulator receivers.

Combiner Improvement

6.17 To this point, microwave receiver noise performance has only been considered for a single receiver. An actual microwave terminal normally will have two or more receivers with the capability to contribute to the baseband output of the terminal. The outputs of the various individual receivers are added (combined) to produce a baseband signal which, if the combiner functions properly, will have superior signal to noise performance when compared to the performance of a single receiver. If the combiners do not function properly, baseband signal level stability and impulse noise performance can be seriously degraded. In general, three types of combining are used: switching (selection), equal gain, and maximal ratio (ratio squared) gain. The effect of combiners is a complicated subject. As Brennan mentions, the relative performance of combiners depends on the type of RSL fading the microwave terminal experiences. There are, however, two cases of special interest. The idealized situation for a LOS terminal is for all receivers to see a constant, unfading RSL which is the same power level at all receivers. For post-detection (baseband) combining, the selection combiner will give no improvement over a single receiver. The equal gain and variable gain combiners will have exactly the same signal to noise improvement. In either case, the improvement will be ten times the common logarithm of the number of receivers used for combining. Unfortunately, this idealized situation is seldom realized in practice. Diversity engineering sometimes intentionally makes the RSL of one receiver different than that of the other receiver. Also, different waveguide run lengths and waveguide hybrid losses can also invalidate the idealization. On the other end of the scale, the TROPO receivers generally experience randomly fading receive signals. If the receivers all have the same average RSL and if the RSL are all Rayleigh amplitude distributed, then the combiner signal to noise improvement will be, on the average, the values that Brennan has predicted. Unfortunately, in real life, for various reasons, the average RSLs of the various receivers is often different by a few dB. Average combiner noise performance under practical conditions is difficult to predict accurately without sophisticated parameter measurement of the actual radio path.

6.18 Combiners serve to improve the baseband signal to noise performance of a microwave terminal. Since the signal out of the combined baseband must be the same level as the signal that was applied to the transmit baseband, the combiner will reduce the noise at the combined baseband when compared to the noise out of a single receiver. As a rough estimate of the actual slot noise that will appear at the combined baseband of

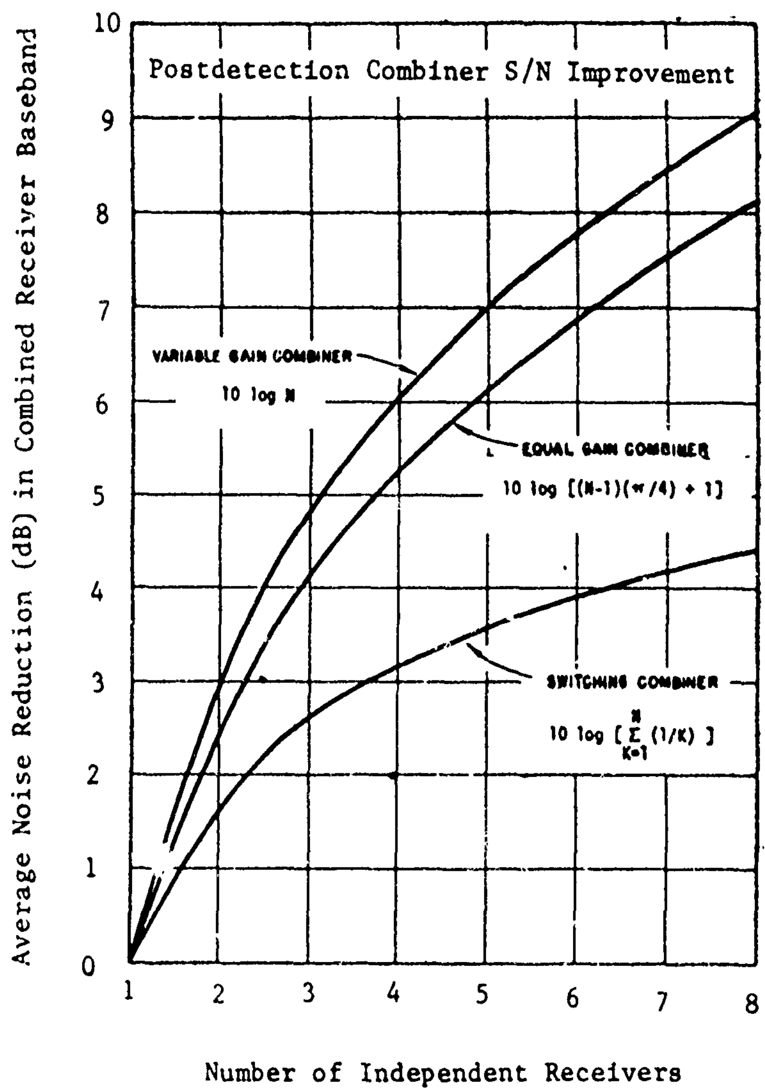
a microwave terminal, the curves Brennan developed can be used. If the terminal is LOS and the RSLs are stable and equal, the following conditions are approximately correct. For selection combining, the noise at the combined baseband will be exactly the same as for a single receiver. The baseband noise can be read directly from the quieting curve of a single receiver. If the terminal uses equal gain or ratio squared combining, the noise at the combined baseband will be less than for a single receiver by the amount shown on the following chart for variable gain combining. The slot noise at combined baseband will be the slot noise read from the quieting curve minus the dB valve read from the Brennan chart. For a TROPO terminal, if the RSLs are the same and all are randomly fading, then the noise improvement will be found by using the curve on the Brennon chart which applies to the type of combining appropriate. The slot noise at the combined baseband will be the slot noise value read from the quieting curve minus the dB noise improvement factor determined from the Brennan curve.

Noise weighting

6.19 When noise measurements are made on telephone circuits, various weighting filters are used to measure the noise. If a 3 kHz flat filter is used to measure the noise, the slot noise previously determined from the quieting curve can be used directly. If a C-Message or Psophometric (CCITT) filter is used, a correction factor must be algebraicly subtracted from the previously determined noise. This subtraction accounts for the noise power lost by the frequency characteristic of the filter. As Tant mentions, the theoretical factor for C-Message is 1.5 dB and is 2.5 for Psophometric weighting. The use of 2.0 dB as the factor for either weighting characteristic is recommended.

Cable Noise

6.20 Cables connect the wideband M/W terminal with multiplexers. The cables generally are located near other signal and power cables. The other cables may induce crosstalk noise. The cables may have noise induced by radio frequency interference. Occasionally, baseband noise will be introduced by the M/W terminal itself. Cable induced noise is difficult to determine without direct measurement. With LOS M/W systems, however, the combined M/W terminal noise and cable induced noise can be measured by terminating the transmit cable at the multiplexer and measuring the cable noise at the multiplexer end of the receive baseband cable using a frequency selective voltmeter. Noise measurements can then be taken with the transmitter baseband amplifier input terminated and the frequency selective voltmeter located directly at the combined terminal receive baseband. Using the two noise measurements and the noise addition/subtraction curves, the cable noise can be obtained.



(after Brennan)

Combiner Improvement Curves

Figure 60

Examples

6.21 To show a method of comparing theoretical quieting curves with actual quieting curve data, two examples will be shown.

LOS M/W Receiver (without de-emphasis):

6.22 The following page lists various slot noise values for various RSLs. The values were taken in dBm at a -15 TLP. The receiver had no de-emphasis. Therefore, to convert the measured valves to dBmØ values, +15 dB must be added to each value. These valves are listed on the following page.

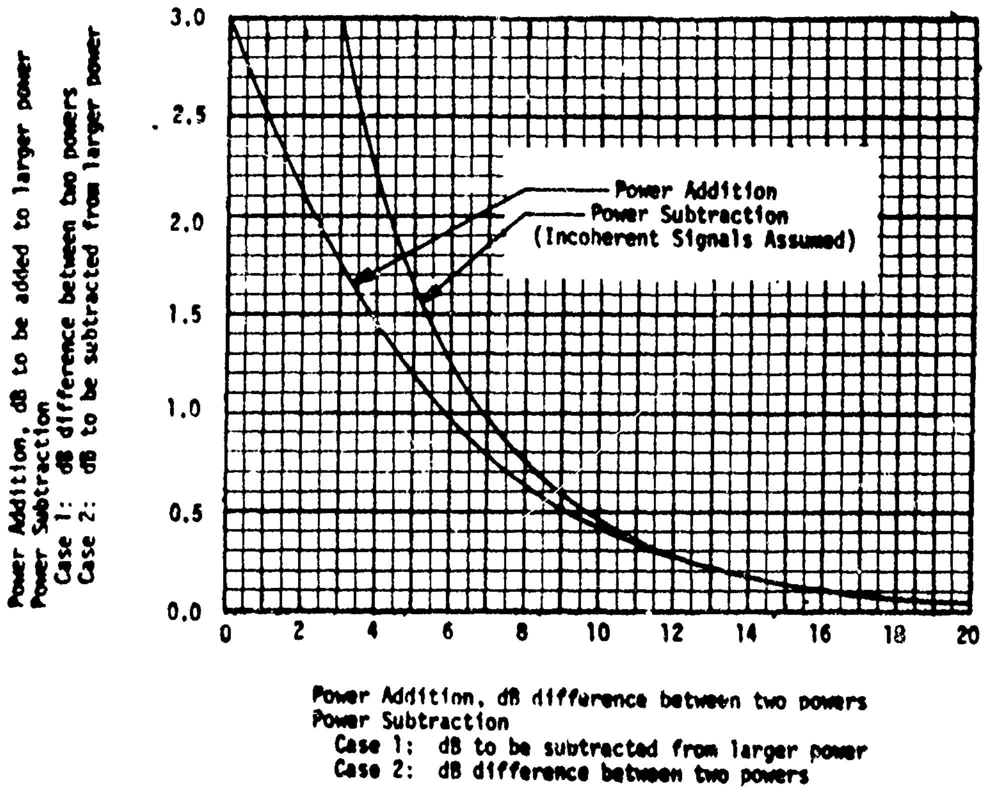
6.23 Following the noise slot measurements, a page is given which lists baseband test tone power level versus various RSLs. The test tone level was measured using a 3.1 kHz wide frequency selective voltmeter (FSV). At low RSLs the noise in the FSV slot was significant. Therefore, what the FSV measured was test tone level plus noise. To compensate for the noise, after measuring the test tone plus noise (TTL + N), the test tone was removed from the baseband of the transmitter and the slot noise (N) was measured at the same RSL. Using the dB subtraction curve of the power addition/subtraction curves the slot noise was subtracted from the test tone level plus noise measurement. This value was listed as the test tone level (TTL). The preceeding values have been plotted on the following graph.

TROPO M/W Receiver (with CCIR de-emphasis):

6.24 The following page lists various receiver baseband slot noise values for various RSLs. The values were taken at a -10 dB TLP. The receiver had CCIR de-emphasis. To convert the measured valves to dBmØ, +10 dB plus the pre-emphasis value (or minus the de-emphasis value) must be added to each valve to arrive at a dBmØ slot noise corrected for de-emphasis.

Slot Frequency (kHz)	TLP (dB)	De-emphasis (dB)	Correction Factor (dB)
308.7	-10	-4.7	+14.7
154.4	-10	+0.8	+ 9.2
77.2	-10	+3.5	+ 6.5
30.9	-10	+4.3	+ 5.7

De-emphasis values taken from actual measurement of the receiver are listed on the next page. The values were taken by running transmitter to receiver baseband frequency response with pre-emphasis strapped out.



Power Addition/Subtraction Curves

Figure 61

FM RECEIVER NOISE QUIETING

[ ] PRELIMINARY DATA		[ ] FINAL DATA	DATE
IF 3dB BW [illegible] MHz	STATION LOS	RECEIVER AN/FRC-157	TEST ENGR / TECH
REF TEST TONE LEVEL 15.0 dBm(0dBm0)		SLOT WIDTH OF F.S. VOLTMETER 3.1 KHz	SLOT WIDTH CORRECTION TO 3.1 KHz 0.0 dB

RSL -dBm	AGC/IF SIGNAL VOLTS (DC) AGC	AGC/IF SIGNAL VOLTS (DC) IF	COMBINER VOLTS (DC)	C/N dB	NOISE POWER IN BASEBAND SLOT (-dBm) — BASEBAND SLOT FREQUENCY f in KHz: MEAS -dBm	CORR TO 3.1 KHz	-dBm0	MEAS -dBm	CORR TO 3.1 KHz	-dBm0	MEAS -dBm	CORR TO 3.1 KHz	-dBm0
						f f/B	2500 0.1	1250 0.05	625 0.025	250 0.01	125 0.005	62.5 0.0025	25 0.001
120				-28			21.5	21	21	21	21	21	21
110				-18			21.5	21.5	21	21	21	21	21
100				-8			22	22	21.5	21.5	21.5	21.5	21.5
98				-6		-dBm	22.5	22	22	22	22	22	22
96				-4		→	23	23	22.5	22.5	22.5	22.5	22.5
94				-2			24	23.5	23.5	23	23.5	23.5	23.5
92				0			25.5	25	25	24.5	24.5	24.5	25
91				+1			26.5	26	26	25.5	25.5	25.5	26
90				+2			27	27	27	26.5	26.5	27	27
89				+3			28.5	28.5	28.5	28	28	28.5	28.5
88				+4			29.5	30	30	29.5	30	30	30
87				+5			32	32.5	33	32.5	32.5	33	33
86				+6			33	34.5	35	35	35	35	35.5
85				+7			35	37	38	38	38	38.5	39
84				+8			36.5	39.5	41.5	41.5	42	42.5	43
83				+9			37.5	42.5	45.5	46.5	47	47	47.5
82				+10			39	44	48	50.5	51	52	52
81				+11			40	45.5	50.5	54.5	56	57	57.5
80				+12			41	46.5	52	57	59.5	60.5	61
79				+13			42	47.5	53	59	62	63.5	64
78				+14			42.5	48	54	60.5	64	65.5	66.5
77				+15			44	49	55	62	66	68	69
76				+16			45	50	56	63	67	69.5	71
75				+17			46	51	57	64	68.5	71.5	73
74				+18			47	52	58	65.5	70	73	75
72				+20			48.5	54.5	60	67.5	73	76.5	78
70				+22			50.5	56	62	69.5	75	79	81.5
68				+24			52.5	58	64	71.5	77.5	81.5	84
66				+26			54.5	60	66	74	79.5	84	86.5
64				+28			56.5	62	68	76	81.5	86	88.5
62				+30			58.5	64.5	70	78	83.5	88	84.5
60				+32			60.5	66.5	72	80	85.5	89	90.5
55				+37			65.5	71	77	84.5	89.5	91.5	91
50				+42			71	76.5	82	89	92.5	92.5	91.5
40				+52			80.5	86	90.5	94	94.5	93	91.5
30				+62			89.5	93	95	95	94.5	93	91.5
20				+72			93	95.5	95.5	95	95	93	91.5
10				+82			95	95.5	95.5	95	95	93	91.5

20 dB QUIETING RSL			FM THRESHOLD RSL		
BASEBAND SLOT FREQUENCY			BASEBAND SLOT FREQUENCY		
KHz	KHz	KHz	KHz	KHz	KHz
dBm	dBm	dBm	dBm	dBm	dBm

Sample Quieting Curve Data

Table 22

FM RECEIVER NOISE QUIETING

PRELIMINARY DATA		FINAL DATA	DATE
IF 3dB BW 25 MHz	STATION LOS	RECEIVER AN/FRC-157	TEST ENGR / TECH
REF TEST TONE LEVEL -15.0 dBm(0dBm0)		SLOT WIDTH OF P.B. VOLTMETER 3.1 KHz	SLOT WIDTH CORRECTION TO 3.1 KHz 0.0 dB

RSL -dBm	AGC/IF SIGNAL VOLTS (DC) AGC	AGC/IF SIGNAL VOLTS (DC) IF	COMBINER VOLTS (DC)	C/N dB	NOISE POWER IN BASEBAND SLOT (-dBm0) — BASEBAND SLOT FREQUENCY f in KHz: MEAS dBm	CORR TO 1.1 KHz	-dBm0	f in KHz: MEAS -dBm	CORR TO 1.1 KHz	-dBm0	f in KHz: MEAS -dBm	CORR TO 1.1 KHz	-dBm0
						f f/B	2500 0.1	1250 0.05	625 0.025	250 0.01	125 0.005	62.5 0.0025	25 0.001
120				-28			6.5	6.0	6.0	6.0	6.0	6.0	6.0
110				-18			6.5	6.5	6.0	6.0	6.0	6.0	6.0
100				-8			7	7	6.5	6.5	6.5	6.5	6.5
98				-6		-dBm0	7.5	7	7	7	7	7	7
96				-4		→	8	8	7.5	7.5	7.5	7.5	7.5
94				-2			9	8.5	8.5	8	8.5	8.5	8.5
92				0			10.5	10	10	9.5	9.5	9.5	10
91				+1			11.5	11	11	10.5	10.5	10.5	11
90				+2			12	12	12	11.5	11.5	12	12
89				+3			13.5	13.5	13.5	13	13	13.5	13.5
88				+4			14.5	15	15	14.5	15	15	15
87				+5			17	17.5	18	17.5	17.5	18	18
86				+6			18	19.5	20	20	20	20	20.5
85				+7			20	22	23	23	23	23.5	24
84				+8			21.5	24.5	26.5	26.5	27	27.5	28
83				+9			22.5	27.5	30.5	31.5	32	32	32.5
82				+10			24	29	33	35.5	36	37	37
81				+11			25	30.5	35.5	39.5	41	42	42.5
80				+12			26	31.5	37	42	44.5	45.5	46
79				+13			27	32.5	38	44	47	48.5	49
78				+14			27.5	33	39	45.5	49	50.5	51.5
77				+15			29	34	40	47	51	53	54
76				+16			30	35	41	48	52	54.5	56
75				+17			31	36	42	49	53.5	56.5	58
74				+18			32	37	43	50.5	55	58	60
72				+20			33.5	39.5	45	52.5	58	61.5	63
70				+22			35.5	41	47	54.5	60	64	66
68				+24			37.5	43	49	56.5	62.5	66.5	[illegible]
66				+26			39.5	45	51	59	64.5	69	71.5
64				+28			41.5	47	53	61	66.5	71	73.5
62				+30			43.5	49.5	55	63	68.5	73	74.5
60				+32			45.5	51.5	57	65	70.5	74	75.5
55				+37			50.5	56	62	69.5	74.5	76.5	76
50				+42			56	61.5	67	74	77.5	77.5	76.5
40				+52			65.5	7	75.5	79	79.5	78	76.5
30				+62			74.5	78	80	80	79.5	78	76.5
20				+72			78	80.5	80.5	80	80	78	76.5
10				+82			80	80.5	80.5	80	80	78	76.5

20 dB QUIETING RSL — BASEBAND SLOT FREQUENCY			FM THRESHOLD RSL — BASEBAND SLOT FREQUENCY		
2500 KHz	250 KHz	250 KHz	2500 KHz	1250 KHz	250 KHz
-80 dBm	-83.5 dBm	-84 dBm	-85 dBm	-84 dBm	-79 dBm

Sample Quieting Curve Ds

Table 23

FM RECEIVER ~~NOISE QUIETING~~ Test Tone Suppression

[ ] PRELIMINARY DATA [ ] FINAL DATA

DATE		
STATION: IOS	RECEIVER: AN/FRC-157	TEST ENGR / TECH
REF. TEST TONE LEVEL: dBm(OdBm0)	SLOT WIDTH OF NOISE VOLTMETER: KHz	dB

RSL -dBm	AGC/IF SIGNAL VOLTS (DC) AGC	IF	COMBINER VOLTS (DC)	C/N dB	MEAS -dBm	CORR TO 3 KHz	-dBm0	MEAS -dBm	CORR TO 3 KHz	-dBm0	MEAS -dBm	CORR TO 3 KHz	-dBm0
						TTL + N	N	TTL					
						(dBu)	(dBu)	(dBu)					
80				+12		0.0	–	0.0		TTL =	Test	Tone	Level
81				+11		-0.1	–	-0.1					
82				+10		-0.2	–	-0.2		N =	Noise		
83				+9		-0.4	–	-0.4					
84				+8		-0.7	–	-0.7		KSL =	Received	Signal	Level
85				+7		-1.2	–	-1.2					
86				+6		-1.6	–	-1.6		dBu =	dB	uncorrected	
87				+5		-2.4	-20.5	-2.4			(relative	dB	value)
88				+4		-3.4	-18.5	-3.5					
89				+3		-4.5	-17.0	-4.7					
90				+2		-5.9	-15.6	-6.4					
91				+1		-7.0	-14.8	-7.8					
92				0		-8.1	-14.1	-9.4					
93				-1		-9.1	-13.5	-11.1					
94				-2		-9.7	-13.1	-12.4					
95				-3		-10.3	-12.6	-4.2					
96				-4		-10.8	-12.5	-[illegible].7					
[illegible]				-5		-10.8	-12.3	-6.1					

20 dB QUIETING RSL			FM THRESHOLD RSL		
BASEBAND SLOT FREQUENCY			BASEBAND SLOT FREQUENCY		
KHz	KHz	KHz	KHz	KHz	KHz
dBm	dBm	dBm	dBm	dBm	dBm

Sample Quieting Curve Data

Table 24

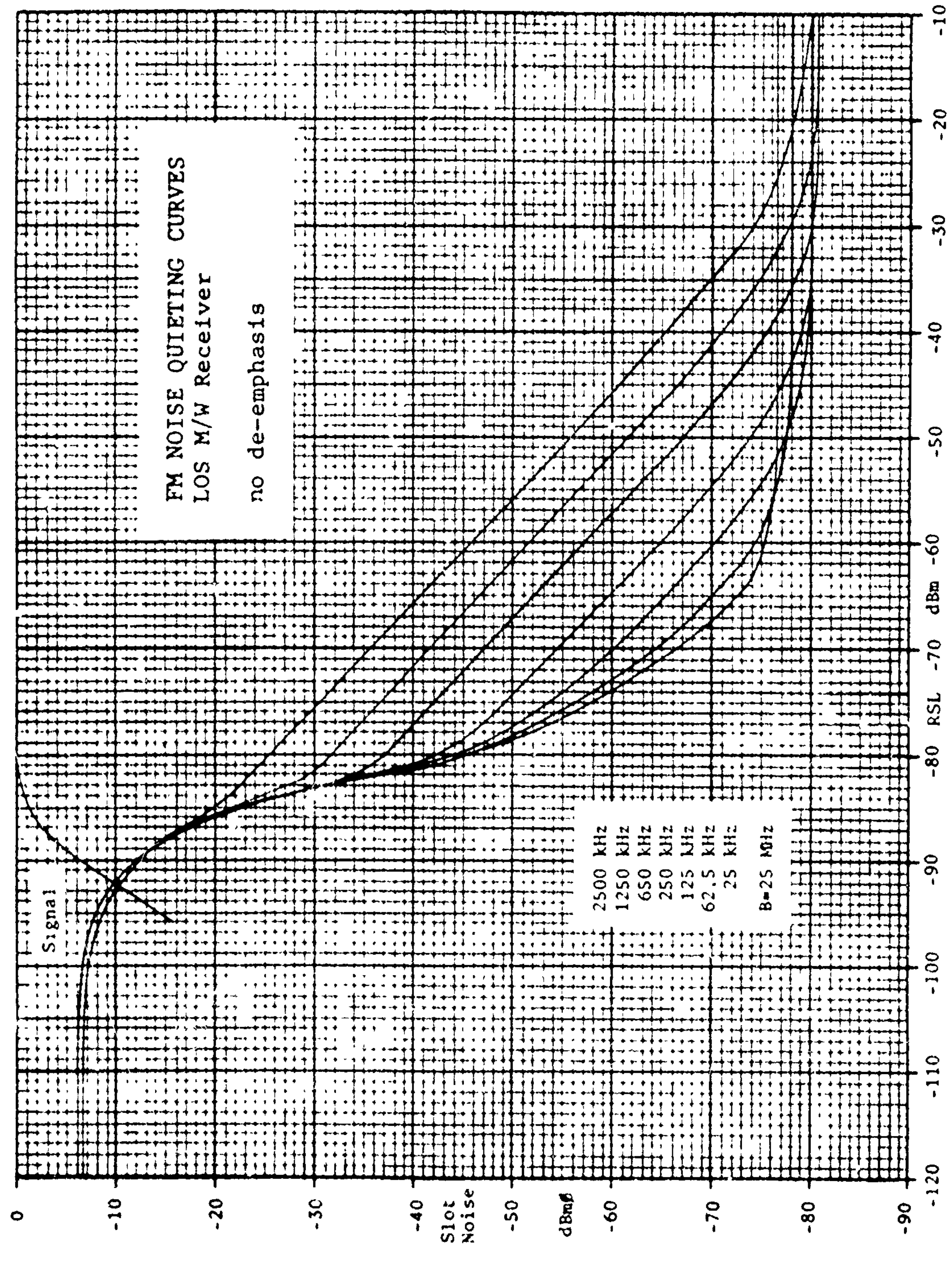

FM NOISE QUIETING CURVES
LOS M/W Receiver
no de-emphasis
Signal
2500 kHz
1250 kHz
650 kHz
250 kHz
125 kHz
62.5 kHz
25 kHz
B=25 MHz
RSL dBm
Slot Noise
dBm0

Figure 62

f(kHz)	f/fmax	De-emphasis (dB) Measured	Theoretical
252	1.0	-3.7	-4.01
227	0.9	-2.6	-3.12
202	0.8	-1.5	-2.07
176	0.7	-0.3	-0.96
151	0.6	+0.2	+0.15
126	0.5	+2.0	+1.19
101	0.4	+2.0	+2.11
75.6	0.3	+3.5	+2.89
50.4	0.2	+4.0	+3.48
25.2	0.1	+4.2	+3.86

f(kHz)	f/B	Measured	Theoretical
309	0.1	-4.7	-4.99
154	0.05	+0.8	+0.01
77.2	0.025	+3.5	+2.85
30.9	0.01	+4.3	+3.79

Tropo M/W Receiver De-emphasis Performance

Table 25

FM RECEIVER NOISE QUIETING

PRELIMINARY DATA [ ] FINAL DATA DATE

IF 3dB BW	STATION	RECEIVER	TEST ENGR	TECH
10.7 MHz	TROPO	AN/FRC-39B		

REF TEST TONE LEVEL	SLOT WIDTH OF FS VOLTMETER	SLOT WIDTH CORRECTION TO 3.1 KHz
10 0 dBm (dBmØ)	3.1 KHz	0.0 dB

NOISE POWER IN BASEBAND SLOT — BASEBAND SLOT FREQUENCY

RSL dBm	AGC/IF SIGNAL VOLTS (DC) AGC	AGC/IF SIGNAL VOLTS (DC) IF	COMBINER VOLTS (DC)	C/N dB	f = ___ KHz MEAS dBm	f = ___ KHz CORR TO 3.1 KHz	f = ___ KHz -dBmØ	f = ___ KHz MEAS -dBm	f = ___ KHz CORR TO 3.1 KHz	f = ___ KHz -dBmØ	f = ___ KHz MEAS -dBm	f = ___ KHz CORR TO 3.1 KHz	f = ___ KHz -dBmØ
					308.7 0.1	154.4 0.05	77.2 0.025	30.9 0.01	f 5/8	308.7 0.1	154.4 0.05	77.2 0.025	30.9 0.01
160				-52	23	17 5	15	14		8 3	8 3	8 5	8.3
120				-12	23	18	15	4 5		8.3	8.8	8.5	8.8
110				-2	25	19 5	17	16		10.3	10.3	10,5	10.3
108				0	26	21	18	17.5	-dBm	11.3	11.8	11.5	11.8
106				+2	28	23	20	19.5	←	13.3	13.8	13.5	13.8
104				+4	30.5	26	23	22.5		15.8	16.8	16.5	16.8
102				+6	35 5	31	28.5	28		20.8	21.8	22	22.3
100				+8	40.5	37 5	35.5	35.5		25.8	28.3	29	29.8
99				+9	43	42	43	44	-dBmØ	28 3	32 8	36.5	38.3
98				+10	44.5	44.5	46.5	50.5	- - ->	29.8	35.3	40	44.8
97				+11	46	46.5	49 5	56	(de-em-	31 3	37 3	43	50.3
96				+12	46.5	47	50.5	57 5	phasis	31.8	37.8	44	51.8
95				+13	47 5	48	51 5	58 5	effect	32 8	38 8	45	52.8
94				+14	48.5	49	52 5	59 5	deleted)	33 8	39 8	46	53.8
93				+15	49 5	50	53.5	61		34 8	40 8	47	55.3
92				+16	50 5	51	54.5	62		35.8	41.8	48	56.3
91				+17	51 5	52	55.5	63		36.8	42.8	49	57.3
90				+18	52.5	53	56.5	64		37.8	43.8	50	58.3
89				+19	53.5	54.5	57.5	65		38 8	44.8	51	59.3
88				+20	54 5	55.5	58.5	66		39.8	45.8	52	60.3
87				+21	55.5	56.5	59.5	67		40.8	46.8	53	61.3
86				+22	56.5	57	60.5	67 5		41.8	47.8	54	61 8
85				+23	57.5	58	61 5	68.5		42.8	48.8	55	62.8
84				+24	58.5	59	62.5	69.5		43.6	49.8	56	63 8
83				+25	59.5	60	63.5	70.5		44.8	50.8	57	64 8
82				+26	60.5	61	64.5	71 5		45 8	51.8	58	65.8
81				+27	61.5	62	65.5	72.5		46.8	52.8	59	66 8
80				+28	62 5	63	66.5	73.5		47 8	53.8	60	67.8
75				+33	67.5	68	71	78		52.8	58 8	64.5	72 8
70				+38	72.5	73	76	83		57 8	63.8	69.5	77.3
65				+43	77	77 5	80 5	85.5		62 3	68.3	74	79.8
60				+48	81 5	82 5	85	87.5		66 8	73.3	78.5	81.8
55				+53	86	86 5	88	88.5		[illegible]	77 3	81.5	82 8
50				+58	89.5	89.5	90	88 5		[illegible]	80.3	83.[illegible]	82 8
45				+63	92	91.5	91	[illegible]		[illegible]	82.3	84.[illegible]	83.3
40				+68	93 5	92.5	91.5	89		78 8	83 3	85	83 3
30				+78	94	93	91.5	89		79 3	83.8	85	83 3
20				+88	94	93	91.5	89		79 3	83.8	85	83 3

20 dB QUIETING RSL			FM THRESHOLD RSL		
BASEBAND SLOT FREQUENCY			BASEBAND SLOT FREQUENCY		
308.7 KHz	154.4 KHz	30.9 KHz	308.7 KHz	154.4 KHz	30.9 KHz
99 dBm	-100 dBm	-100 dBm	-100 dBm	-99 dBm	-98 dBm

Sample Quieting Curve Data

Table 26

FM RECEIVER ~~NOISE QUIETING~~ Test Tone Suppression													
[ ] PRELIMINARY DATA						[ ] FINAL DATA				DATE			
	STATION: TROPO				RECEIVER: AN/FRC-39B					TEST ENGR		TECH	
REF TEST TONE LEVEL dBm(0dBm0)					SLOT WIDTH OF F.S. VOLTMETER KHz					SLOT WIDTH CORRECTION TO 3.1 KHz dB			

RSL -dBm	AGC IF SIGNAL VOLTS DC: AUC	AGC IF SIGNAL VOLTS DC: II	COMBINER VOLTS (DC)	C/N dB	~~NOISE POWER IN BASEBAND SLOT~~ ~~BASEBAND SLOT FREQUENCY~~ KHz: MEAS -dBm	CORR TO 3.1 KHz	-dBm0	KHz: MEAS -dBm	CORR TO 3.1 KHz	-dBm0	KHz: MEAS -dBm	CORR TO 3.1 KHz	-dBm0
						TTL + N	N	TTL					
						(dBu)	(dBu)	(dBu)					
83.6				+24		0.0	-	0.0		TTL =	Test	Tone	Level
93.6				+14		0.0	-	0.0					
97.6				+10		-0.1	-	-0.1		N =	Noise		
99.6				+8		-0.25	-	-0.25					
100.6				+7		-0.4	-	-0.4		RSL =	Received	Signal	Level
101.6				+6		-0.9	-20.0	-0.9					
102.6				+5		-1.3	-18.0	-1.4		dBu =	dB	uncorrected	
103.6				+4		-2.0	-16.2	-2.2			(relative	dB	value)
104.6				+3		-2.6	-14.7	-2.9					
105.6				+2		-3.6	-13.2	-4.1					
106.6				+1		-4.4	-12.2	-5.2					
107.6				0		-5.7	-11.2	-7.1					
108.6				-1		-6.5	-10.4	-8.7					
109.6				-2		-7.1	-10.0	-10.1					
110.6				-3		-7.5	-9.6	-11.5					
111.6				-4		-7.9	-9.2	-13.9					
112.6				-5		-8.0	-8.9	-14.5					

20 dB QUIETING RSL			FM THRESHOLD RSL		
BASEBAND SLOT FREQUENCY			BASEBAND SLOT FREQUENCY		
KHz	KHz	KHz	KHz	KHz	KHz
dBm	dBm	dBm	dBm	dBm	dBm

Sample Quieting Curve Data

Table 27

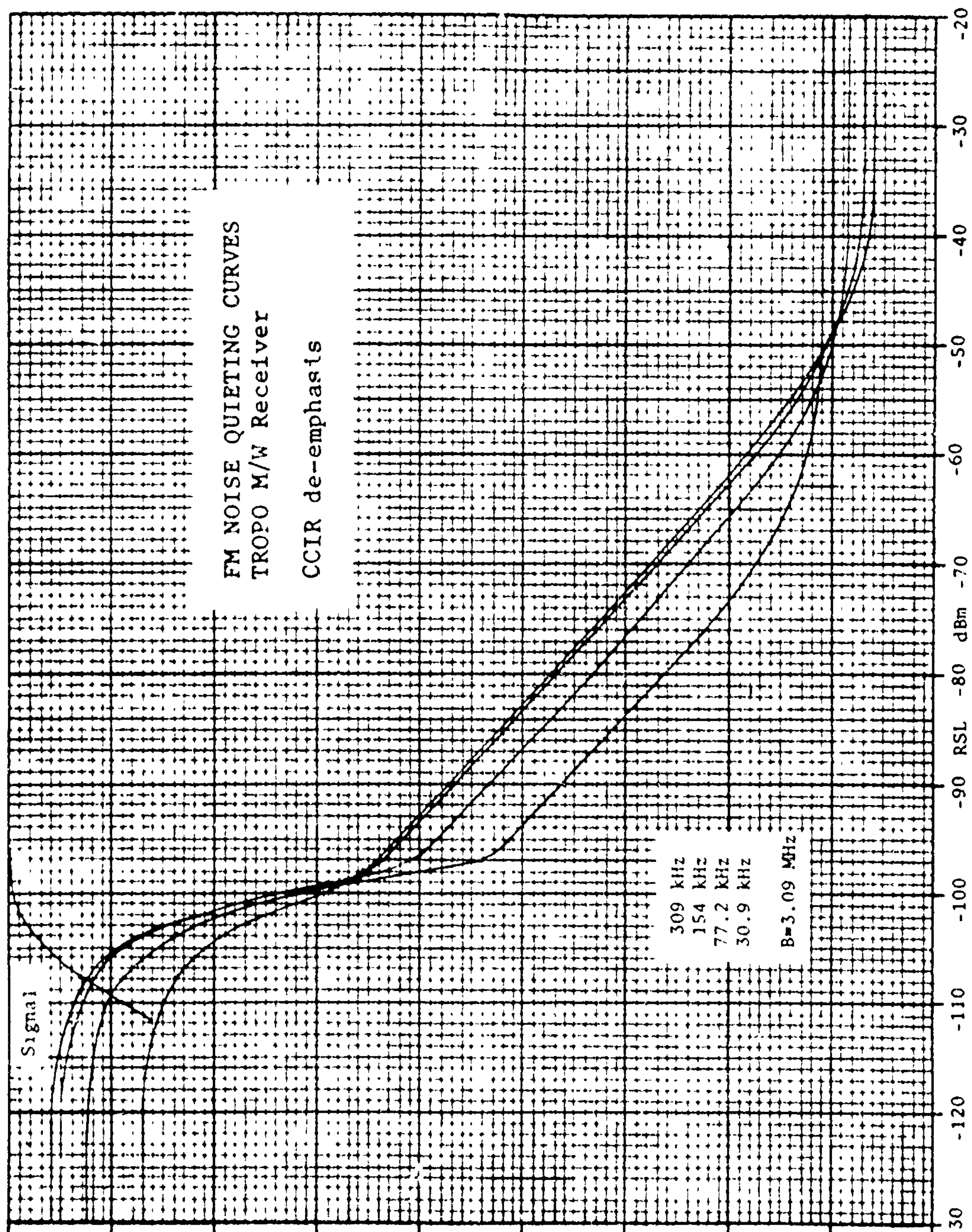

FM NOISE QUIETING CURVES
TROPO M/W Receiver
CCIR de-emphasis
309 kHz
154 kHz
77.2 kHz
30.9 kHz
B=3.09 MHz
Signal
Slot Noise
RSL dBm

Figure 63

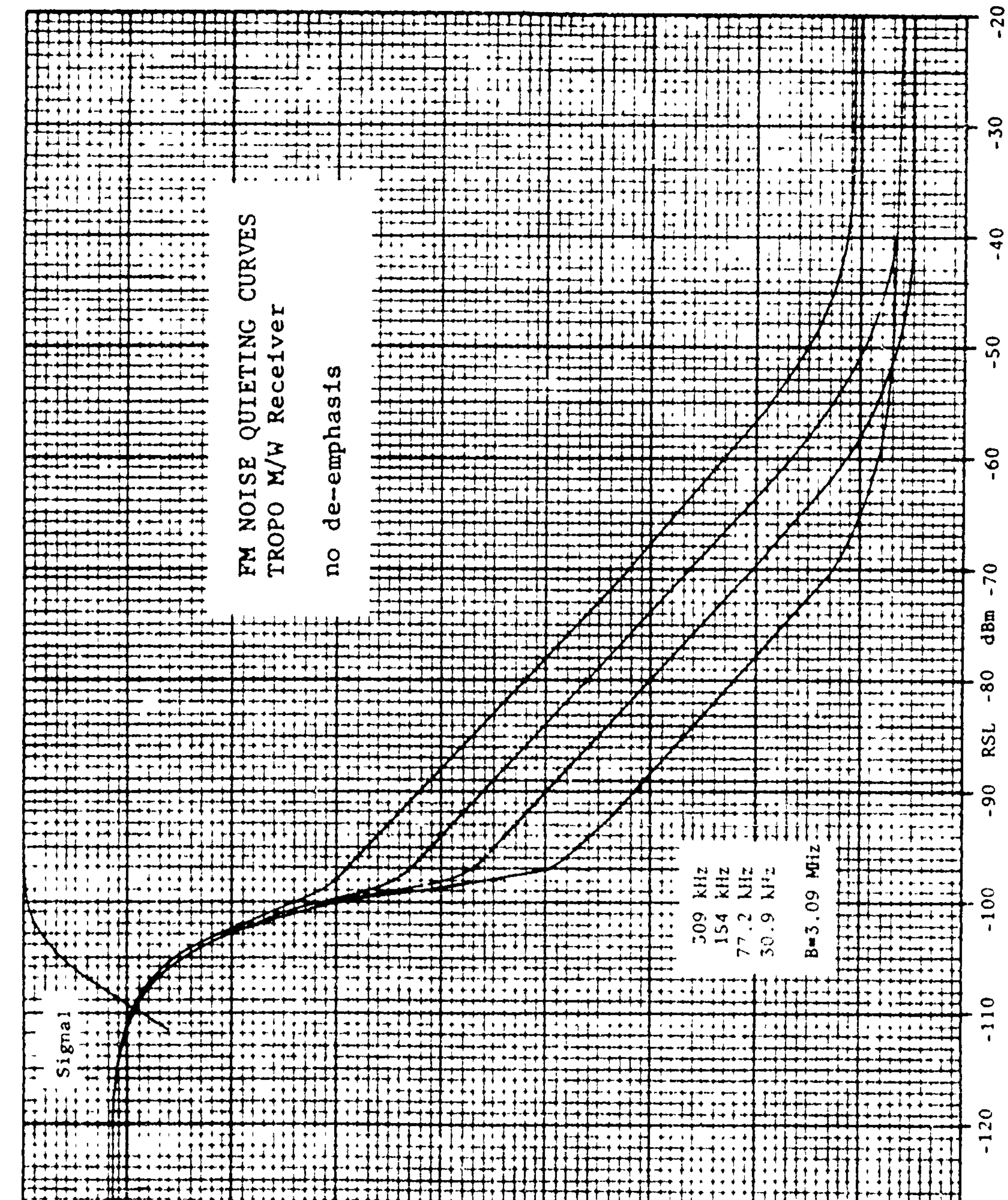
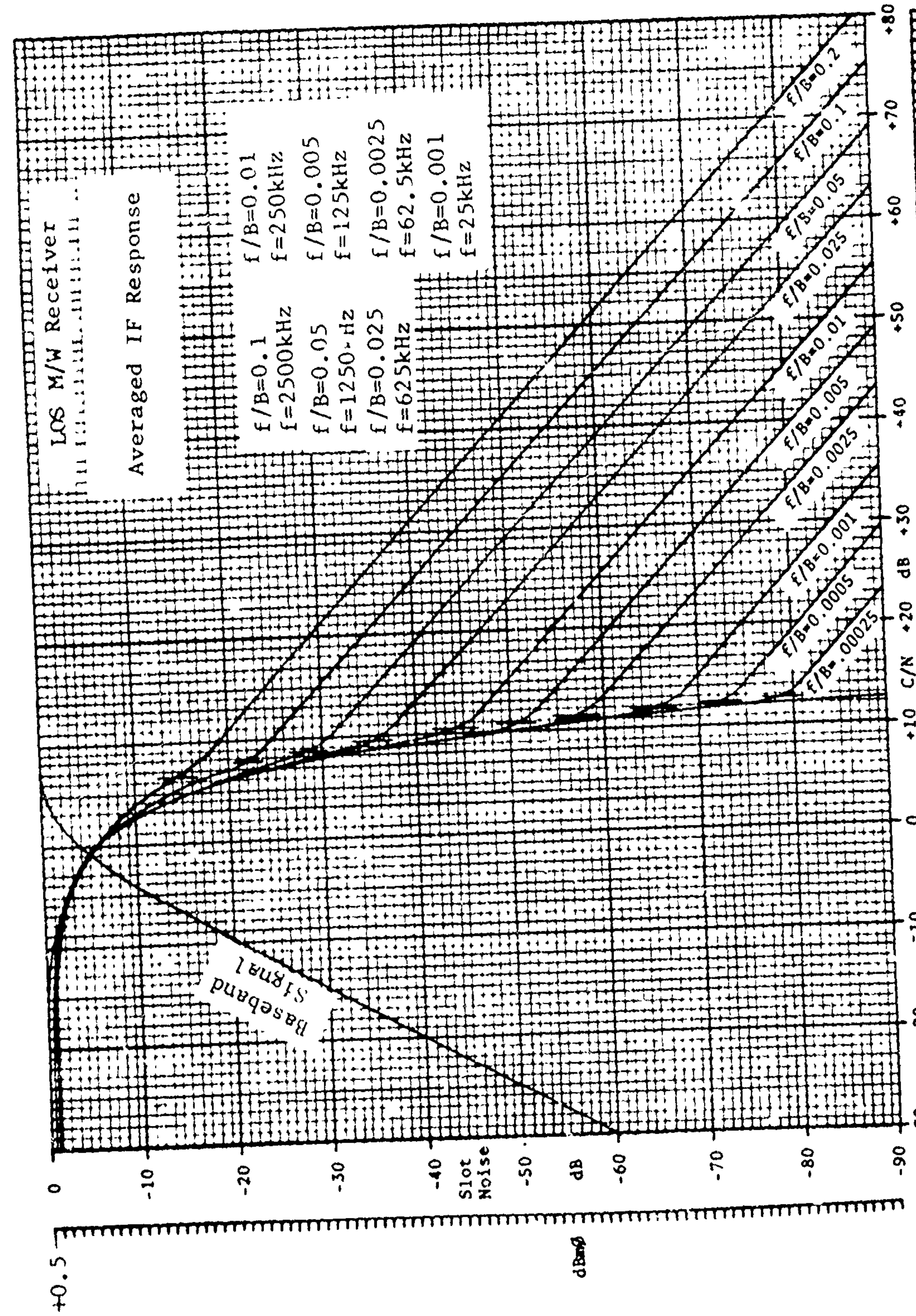
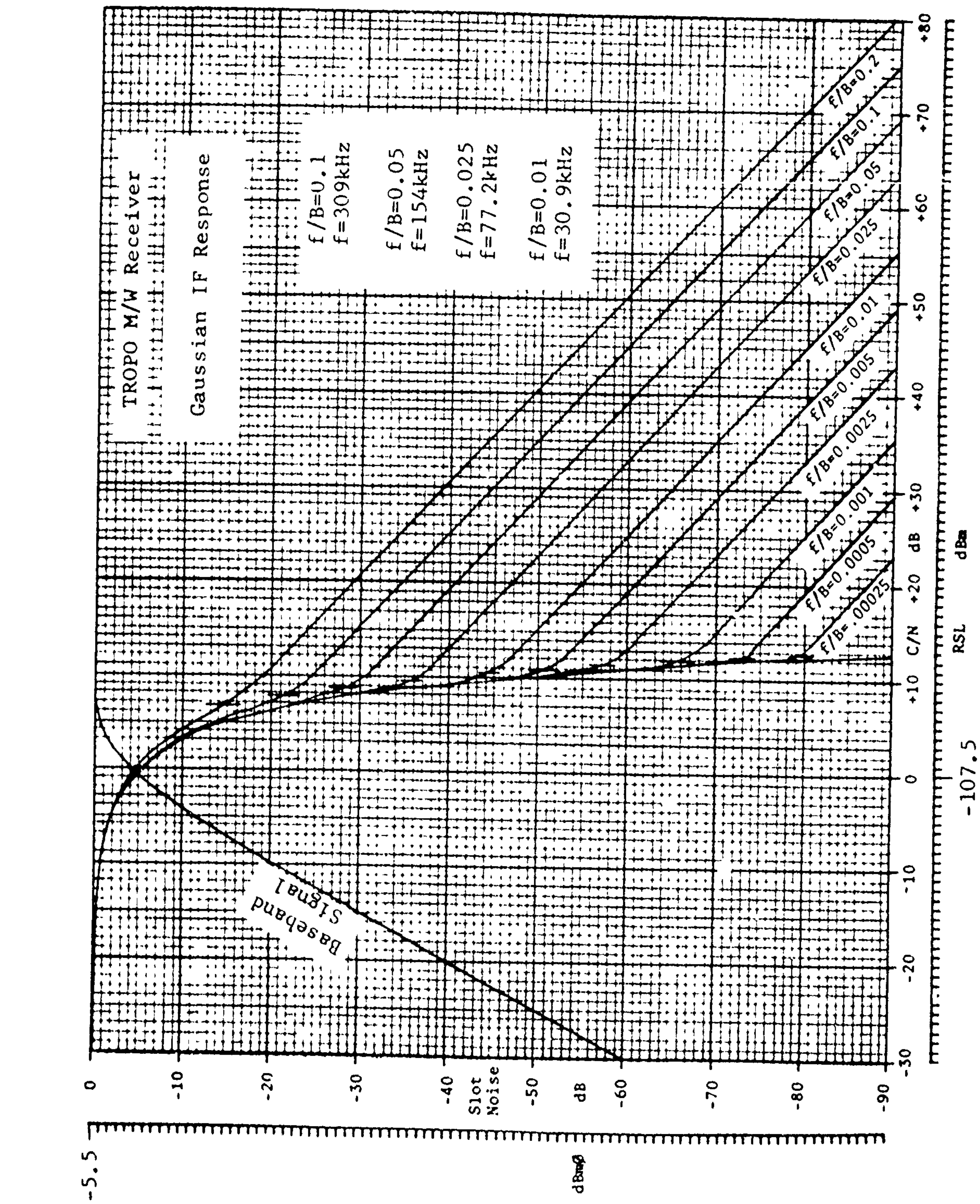

FM NOISE QUIETING CURVES
TROPO M/W Receiver
no de-emphasis
Signal
309 kHz
154 kHz
77.2 kHz
30.9 kHz
B=3.09 MHz
RSL dBm
Slot Noise
dBm0

Figure 64



LOS M/W Receiver
Averaged IF Response
f/B=0.1
f=2500kHz
f/B=0.05
f=1250kHz
f/B=0.025
f=625kHz
f/B=0.01
f=250kHz
f/B=0.005
f=125kHz
f/B=0.0025
f=62.5kHz
f/B=0.001
f=25kHz
f/B=0.2
f/B=0.1
f/B=0.05
f/B=0.025
f/B=0.01
f/B=0.005
f/B=0.0025
f/B=0.001
f/B=0.0005
f/B=.00025
Baseband Signal
Slot Noise
dB
C/N dB
RSL dBm
-92
+0.5
dBm0

Figure 65



TROPO M/W Receiver
Gaussian IF Response
f/B=0.1
f=309kHz
f/B=0.05
f=154kHz
f/B=0.025
f=77.2kHz
f/B=0.01
f=30.9kHz
f/B=0.2
f/B=0.1
f/B=0.05
f/B=0.025
f/B=0.01
f/B=0.005
f/B=0.0025
f/B=0.001
f/B=0.0005
f/B=.00025
Baseband Signal
Slot Noise dB
C/N dB
RSL dBm
-107.5
-5.5
dBm0

Figure 66

6.25 Notice that the measured values are significantly different than the theoretical values. This shows the necessity of checking FM receiver de-emphasis characteristics if accurate comparisons are to be made between actual and theoretical quieting curves.

6.26 Following the noise slot page is a page listing baseband test tone levels versus RSL. Using the same procedure as mentioned for the LOS receiver, values for test tone level were determined from the test tone plus noise measurements. The results have been plotted on the following pages. The first graph shows the quieting curve without de-emphasis correction. The second graph shows the quieting curve after correction for de-emphasis.

6.27 Calculations for using the Averaged RF/IF response chart with the LOS receiver (excluding effects of de-emphasis) follow:

Receiver Specifications:

IF bandwidth: 25 MHz
per channel deviation:: 140 kHz rms
(for 0 dBmØ test tone)
noise figure: 8 dB

Noise Slot Measurement (FSV) Bandwidth: 3.1 kHz

Noise value corresponding to 0 dB slot noise on chart:

$$N = +29.6 - 20 \log \Delta f_{/ch\ rms} + 10 \log B_{IF} + CF$$

$$= +29.6 - 20 \log (140) + 10 \log (25) + 0$$

$$= +29.6 - 42.9 + 14.0 + 0 = +0.7 \text{ dB} = +0.5 \text{ dB}$$

RSL value corresponding to 0 dB C/N:

$$RSL = -114.0 + NF + 10 \log B_{IF}$$

$$= -114.0 + 8.0 + 10 \log (25)$$

$$= -114.0 + 8.0 + 14.0 = -92.0 \text{ dBm}$$

The above data is noted on the following page.

Calculations for using the Gaussian RF/IF response chart with the TROPO receiver (excluding effects of de-emphasis) follow:

Receiver Specifications:

IF bandwidth: 3.09 MHz
per channel deviation: 100 kHz rms
(for OdBmØ test tone)
noise figure: 1.5 dB

Noise Slot Measurement (FSV) Bandwidth: 3.1 kHz

Noise valve corresponding to 0 dB slot noise on chart:

$$N = +29.6 - 20 \log \Delta f_{/ch\ rms} + 10 \log B_{IF} + CF$$

$$= +29.6 - 20 \log (100) + 10 \log (3.09) + 0$$

$$= +29.6 - 40.0 + 4.9 = -5.5 \text{ dBm}Ø$$

RSL value corresponding to OdB C/N:

$$RSL = -114.0 + NF + 10 \log B_{IF}$$

$$= -114.0 + 1.5 + 10 \log (3.09)$$

$$= -114.0 + 1.5 + 4.9 = -107.6 \cong -107.5 \text{ dBm}$$

The above data is noted on the following page.

Note: For ease in comparison with actual quieting curves, it is suggested that the above calculated valves be rounded to the nearest ½ dB.

7. FM Microwave Radio Terminal Equipment Parameters

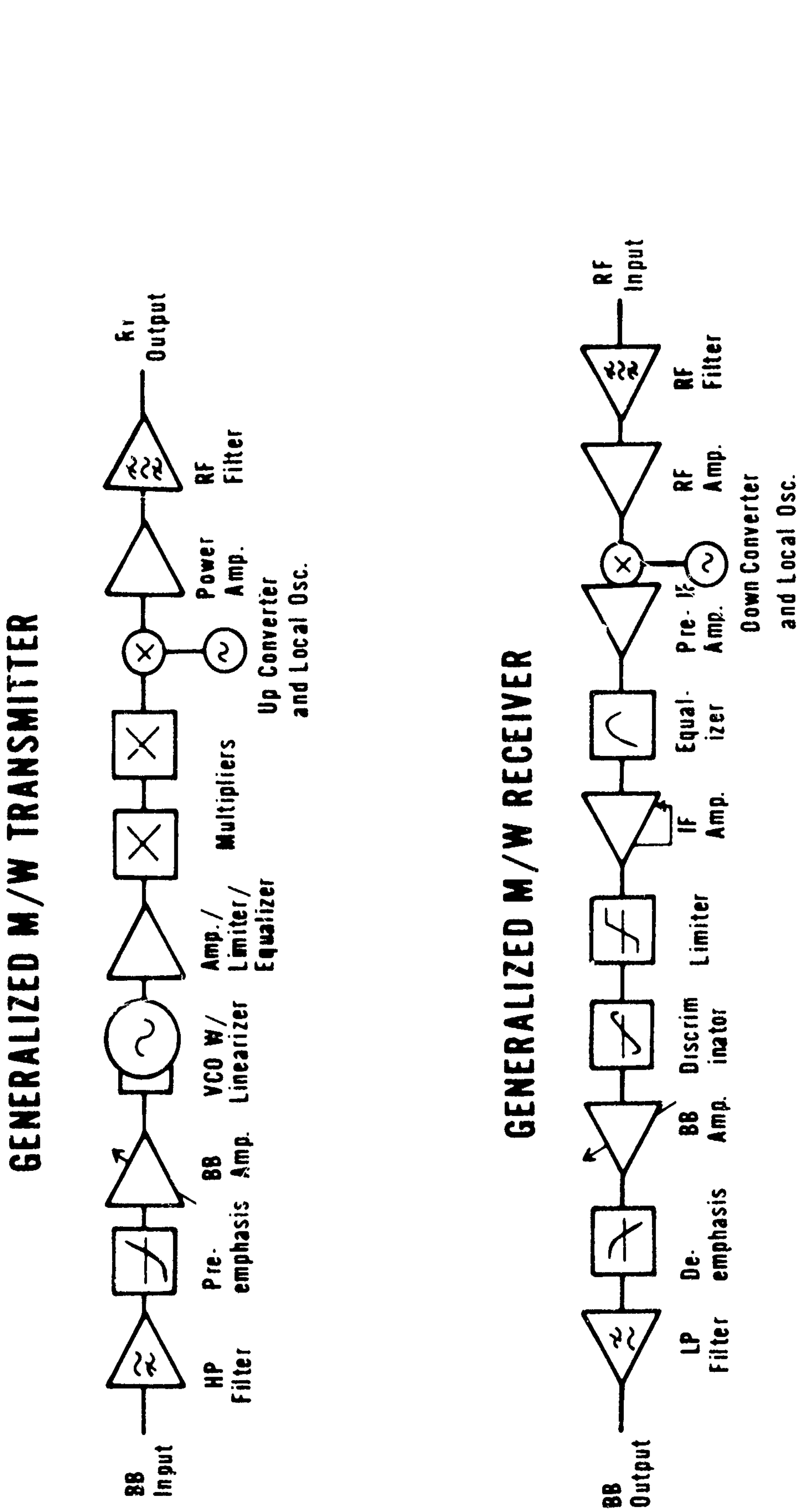
7.1 In order to predict M/W terminal performance, it is necessary to determine a few characteristics of the terminal transmitter and receiver. The determination of these parameters will be discussed in the next few sections. To help keep things in perspective, the following page diagrams a generalized M/W transmitter and receiver. No single transmitter or receiver actually has all the features shown on the diagram. Any given M/W transmitter and receiver can have several of the features on the diagram.

Per Channel RMS Deviation

7.2 One of the parameters that must be known about a microwave system is its sensitivity to deviation of the carrier from its rest or no modulation frequency. At the transmitter we wish to know how far the transmitter is deviated by a given signal. At the receiver we wish to know how much signal is produced by a given frequency deviation of the received carrier. The most common method of specifying these parameters is to specify what is called the "per channel rms deviation" of the radio system. EIA Standard RS-252-A, paras 2.4.2 and 1.15 define "per channel rms deviation" as the rms carrier deviation caused by a 0 dBmØ sine wave test tone. For systems employing emphasis, the deviation must be specified for a specific baseband frequency. By common usage, this frequency is pivot frequency (see the section on emphasis). It is worth mentioning that for an FM system, the frequency deviation of the carrier is directly proportional to the instantaneous voltage of the modulating signal. Therefore, the change in carrier frequency is directly proportional to the change in voltage at the modulator. Since the peak voltage of a sine wave is $\sqrt{2}$ times the rms voltage, the peak deviation of a carrier is $\sqrt{2}$ times the rms deviation (for a modulating sine wave). Also, for an FM system (without emphasis), the per channel rms deviation is constant regardless of frequency. Unfortunatly, this is not true for systems with emphasis.

7.3 We wish to determine the rms deviation caused by a 0 dBmØ sine wave. The most common method of doing this is to take advantage of the properties of the RF spectrum of an FM modulator driven by a sine wave. This method, known as the Crosby null, Bessel null, or carrier dropout method has been described by several sources (e.g., Seal and Rothrock).

7.4 When an FM modulator is driven by a sine wave, the carrier as well as the sidebands disappear (drop out) for certain unique values of a factor called beta. Beta (β) for a sine wave is defined as the ratio of the peak carrier deviation (caused by the sinewave) of the carrier divided by the frequency of the sine wave.



Generalized M/W Transmitter and Receiver

Figure 67

$$\beta = \frac{\Delta f peak}{fm}$$

7.5 We will take advantage of the properties of the carrier and first sideband dropouts. The beta/dropout chart on the following page lists the values of beta for various carrier and first sideband dropouts.

7.6 Let's determine a formula for determining the per channel rms deviation of a transmitter (at its RF output) based on carrier dropout measurements taken with a spectrum analyzer.

7.7 From the beta/dropout chart we know that first carrier dropout occurs when beta is 2.4048.

$$\beta = \frac{\Delta f peak}{fm} = 2.4048$$

or

$$\Delta f peak = 2.4048\ (fm)$$

7.8 We want rms deviation rather than peak deviation. Therefore, we divide by $\sqrt{2}$.

$$\Delta frms = \frac{\Delta f peak}{\sqrt{2}} = \frac{2.4048 \quad fm}{\sqrt{2}} = 1.700\ fm$$

7.9 At this point, we know that a 0 dBmØ sine wave at the input of an FM transmitter without pre-emphasis causes first carrier dropout when observed with a spectrum analyzer at the RF output of the FM transmitter. For this simple situation we would be able to determine the per channel rms deviation by using the preceding formula. Unfortunately, the problem is seldom that simple. Many FM systems have pre-emphasis, 0 dBmØ signals are not always used to dropout the carrier (for various reasons), and dropout is not always observed at the RF output of the radio. Therefore, some correction factors are necessary to account for these practical aspects.

7.10 The first correction factor is for the case when carrier dropout is achieved at a level L (dBmØ) which is not 0 dBmØ. The preceding formula will give us the deviation caused by whatever level that was used to produce it. What is desired, however, is the deviation that would have been produced by a 0 dBmØ test tone. Deviation is directly proportional to voltage. Therefore, the dB difference between L and 0 dBmØ must be converted to a voltage ratio. The convertion factor becomes

Dropout Number	Carrier	First Sideband	Carrier	First Sideband	First Sideband / Carrier
n	β_n	β_n	dB(β_n/β_1)	dB(β_n/β_1)	dB(β_n/β_n)
1	2.4048	3.8317	0.00	0.00	4.05
2	5.5201	7.0156	7.22	5.25	2.08
3	8.6537	10.1735	11.12	8.48	1.41
4	11.7915	13.3237	13.81	10.82	1.06
5	14.9309	16.4706	15.86	12.67	0.85
6	18.0711	19.6159	17.52	14.18	0.71
7	21.2116	22.7601	18.91	15.48	0.61
8	24.3525	25.9037	20.11	16.60	0.54
9	27.4935	29.0468	21.16	17.59	0.48
10	30.6346	32.1897	22.10	18.49	0.43
11	33.7758	35.3323	22.95	19.30	0.39
12	36.9171	38.4748	23.72	20.04	0.36
13	40.0584	41.6171	24.43	20.72	0.33
14	43.1998	44.7593	25.09	21.35	0.31
15	46.3412	47.9015	25.70	21.94	0.29
16	49.4826	51.0435	26.27	22.49	0.27
17	52.6241	54.1856	26.80	23.01	0.25
18	55.7655	57.3275	27.31	23.50	0.24
19	58.9070	60.4695	27.78	23.96	0.23
20	62.0485	63.6114	28.23	24.40	0.22

TABLE OF BETAS AND DROPOUTS

Frequency Modulation with Sinewave Modulation is Assumed.

Beta = Peak Frequency deviation (relative to carrier rest frequency)/modulating frequency = β_n

dB(x) = 20 log (x)

Table 28

$$10^{-L/20}$$

where L is in dBmØ

7.11 Sometimes it is necessary or convenient to measure deviation at a test point prior to transmitter multiplier sections. When the transmit signal is multiplied in frequency, the deviation is multiplied by the same factor. The rms deviation we want is the deviation at the output of the transmitter. Therefore, the deviation derived by the above formula must be multiplied by M, the product of all the multipliers between the test point and the RF output of the radio.

7.12 If the transmitter has pre-emphasis two additional correction factors are required. The level of a sine wave is changed (depending on its frequency) as it passes through a pre-emphasis network. If pre-emphasis was not in the circuit, the level must be correct to account for the level that would occur at pivot frequency (see the section on emphasis) when the pre-emphasis circuit is reinstalled. This level correction is Pp and is the level in dB of a sine wave at pivot frequency with pre-emphasis in to the level of the same sinewave with pre-emphasis out. If carrier dropout is achieved with a frequency other than pivot frequency, an additional level correction factor P(fm) must be added. This factor is the level change (in dB) of the sine wave as it passes through the pre-emphasis network compared to the level change as the sine wave passed through the pre-emphasis network at pivot frequency. This is called pre-emphasis relative to pivot frequency.

Putting all the factors together we get the following formula:

$$\Delta f_{/ch\ rms} = 1.700 \times M \times fm \times \text{antilog}\ ((\ P(fm) + Pp - L)/20)$$

where

$\Delta f_{/ch\ rms}(kHz)$ = per channel rms deviation at the RF output of the radio (for a 0 dBmØ test tone at pivot frequency).

M = composite multiplication factor between the test point at which first carrier dropout is observed and the final RF output port of the radio. For example, if dropout is observed prior to cascaded X 2 and X 3 multipliers followed by an up converter (there is no change in deviation due to an up/down converter), then M = 3x2 = 6.

fm(kHz) = frequency of test tone used to achieve first carrier dropout.

L(dBmØ) = level of test tone used to achieve first carrier dropout.

$X(dBm\emptyset) = Y(dBm) - TLP(dB)$

$P(fm)(dB)$ = pre-emphasis effect at frequency fm relative to pre-emphasis effect at pivot frequency. This factor is 0.0 if pre-emphasis is strapped out when dropout is observed or if the system does not have pre/de-emphasis.

$Pp(dB)$ = pre-emphasis effect at pivot frequency relative to test tone level with pre-emphasis strapped out. This factor is 0.0 for all modern (CCIR/EIA) preemphasis networks, but is about 8.0 for the older time constant and 4 dB/octave networks. This factor is 0.0 if no pre/de-emphasis is used in the system. This factor is 0.0 if dropout is achieved with pre-emphasis in the circuit.

7.13 Sometimes it is necessary to reverse the above process. The per channel rms deviation is known but a frequency and level must be determined which will cause first carrier dropout.

7.14 Given a per channel rms deviation and a frequency, first carrier dropout will occur for the following sinewave level:

$$L = P(fm) + Pp + 20 \log (1.700 \times fm) - 20 \log \frac{\Delta f_{/ch\ rms}}{M}$$

Given a per channel rms deviation and a level, first carrier dropout will occur for the following frequency.

$$fm = \frac{\Delta f_{/ch\ rms}}{1.700\ (M)\ \text{antilog}\ ((\ P(fm) + Pp - L)/20)}$$

Example Calculations:

7.15 Consider first carrier dropout using a 83.333 kHz test tone. Assume RF per channel RMS deviation of 140 kHz and no pre/de-emphasis.

$fm = 83.333$ kHz

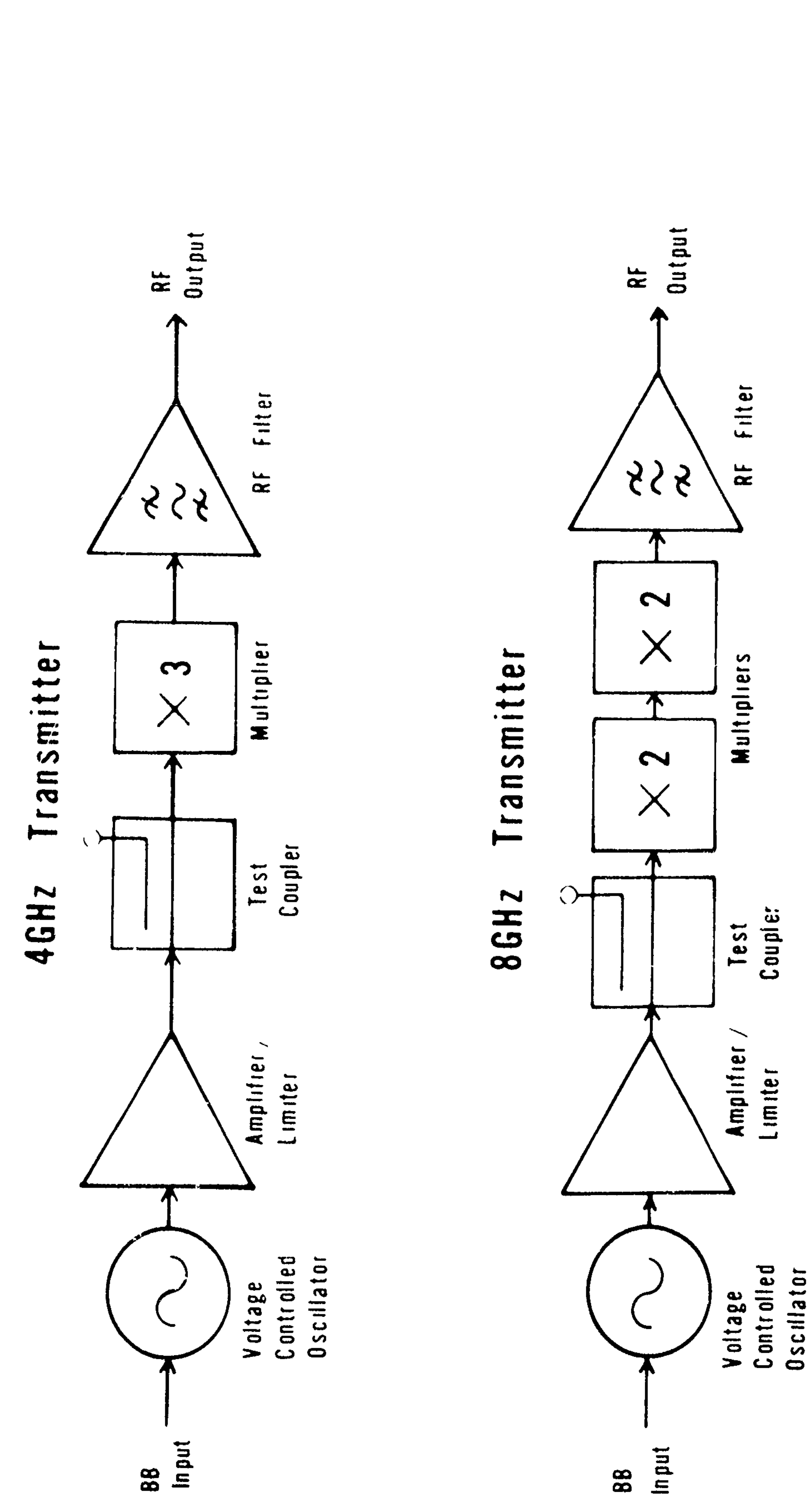
$\Delta f_{/ch\ rms} = 140$ kHz

$P(fm) = Pp = 0.0$ (no pre-emphasis)

7.16 With reference to the hypothetical FM M/W transmitters diagramed on the next page, the various levels necessary to cause first carrier dropout are determined:

Assume dropout is measured at the RF output port (M=1):

$$L = 0 + 0 + 20 \log (1.7 \times 83.333) - 20 \log (140/1) = +0.11\ dBm\emptyset$$



Example Transmitter Configurations

Figure 68

Assume dropout is measured at the Test Coupler output (consider 4GHz version, M=3):

$$L = 0 + 0 + 20 \log (1.7x83.333) - 20 \log (140/3) = +9.65 \text{ dBm}0$$

7.17 Assume dropout is measured at the Test Coupler output (consider 8 GHz version, M=2x2=4):

$$L = 0 + 0 + 20 \log (1.7x83.333) - 20 \log (140/4) = +12.15 \text{ dBm}0$$

7.18 Notice that for a radio specification of 140 kHz per channel RMS RF deviation and given a 83.333 kHz baseband test tone, first carrier dropout can be achieved with one of three possible test tone levels.

7.19 The carrier dropout method is quite useful in determining per channel rms deviation. However, this method is not perfect. When this method is used, a spectrum analyzer is used to monitor the carrier null. There are many carrier nulls possible, however. The immediate question is "How do you know which null has occurred?" One way is to take advantage of the relationship between the power level required to achieve first carrier dropout and the power required to achieve first sideband dropout. The right hand column of the previous beta/dropout chart lists the change in modulator drive required to go from a given order carrier dropout to a given order first sideband dropout. To go from first carrier dropout to first first sideband dropout requires an increase of almost exactly 4 dB. All the other dropouts require less than 4 dB to go from carrier to first sideband dropout. Therefore, if increasing the drive to the FM modulator 4 dB causes the carrier dropout to become a first carrier dropout, then the previous carrier dropout was the first carrier dropout.

7.20 If changing the drive level exactly 4 dB is not convenient, a couple of other methods are available. If a spectrum analyzer with calibrated vertical (power) scale is available, the power levels of the first few sidebands around the carrier frequency can be determined. The following page lists the power levels for the first five sidebands at the first four carrier and first sideband dropouts. The 0 dB reference power is the power level of the unmodulated carrier.

7.21 Even without a calibrated spectrum analyzer, the dropout can be determined. Each carrier dropout causes a distinctive overall shape to the envelope of the sideband power levels. This shape is independent of the frequency of the test tone used to produce the dropout. The following pages show the spectral display for the first ten carrier dropouts. With these pictures and a little practice, the correct dropout can be determined by inspection.

7.22 There are a couple of other problems to watch out for when doing carrier dropout. It is possible to achieve carrier dropout even if a modulation amplifier or voltage controlled oscillator/klystron is not linear. The two most common nonlinearities are clipping the signal

BETA	CARRIER	SIDEBAND 1	SIDEBAND 2	SIDEBAND 3	SIDEBAND 4	SIDEBAND 5
2.40	-97.54	-5.69	-7.30	-14.02	-23.78	-35.71
5.52	-102.44	-9.36	-18.18	-12.01	-8.05	-9.82
8.65	-99.[illegible]4	-11.33	-24.05	-12.31	-12.73	-30.73
11.79	-88.01	-12.67	-28.09	-13.19	-16.43	-18.66

Carrier Dropouts

BETA	CARRIER	SIDEBAND 1	SIDEBAND 2	SIDEBAND 3	SIDEBAND 4	SIDEBAND 5
3.83	-7.90	-112.37	-7.90	-7.53	-11.85	-18.92
7.02	-10.45	-107.66	-10.45	-15.33	-16.26	-9.21
10.17	-12.05	-100.51	-12.05	-20.16	-14.34	-12.08
13.32	-13.21	-80.99	-13.21	-23.66	-14.48	-14.95

First Sideband Dropouts

Dropout Power Levels (dB)

Table 29

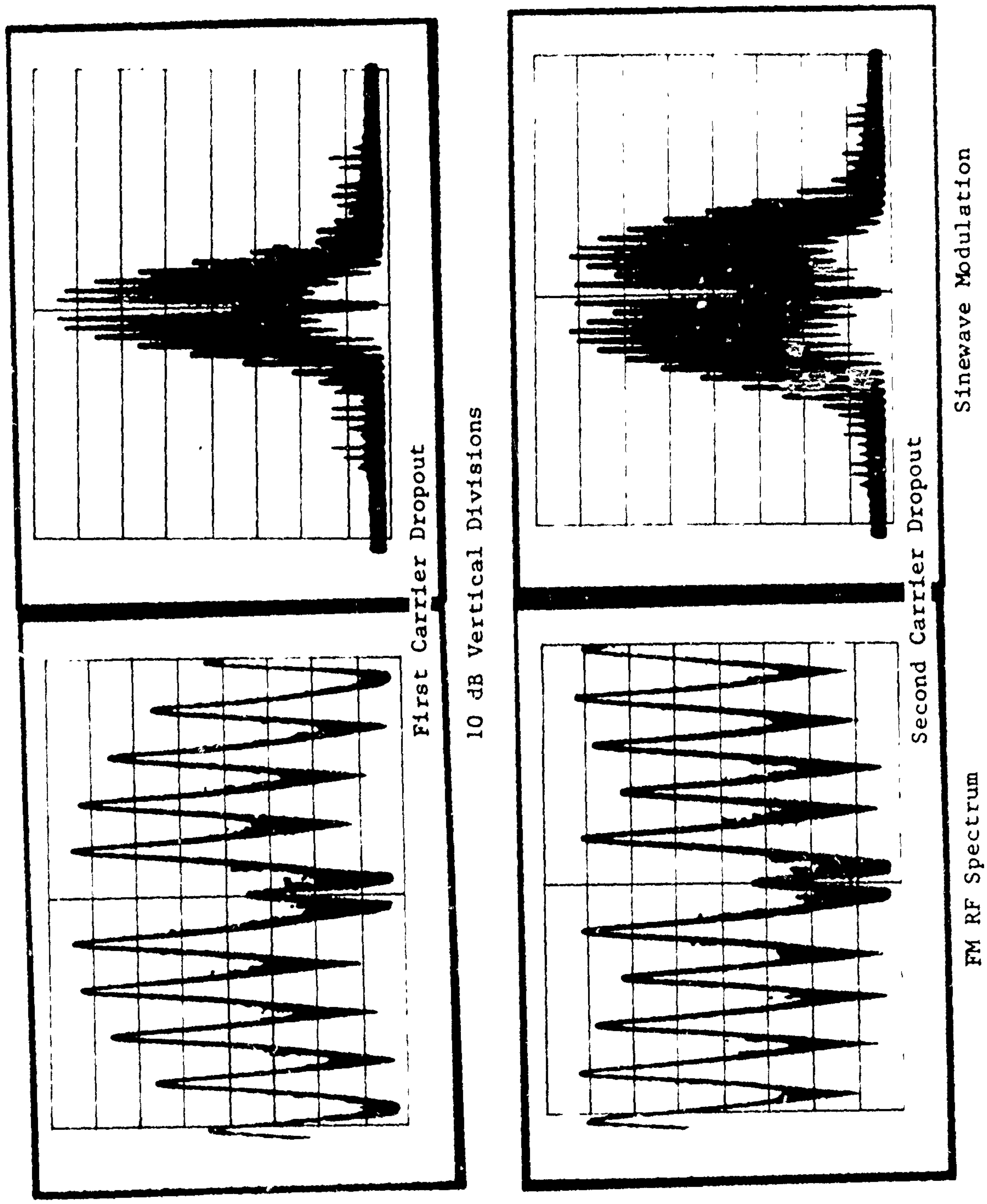

First Carrier Dropout
Second Carrier Dropout
10 dB Vertical Divisions
Sinewave Modulation
FM RF Spectrum

Figure 69

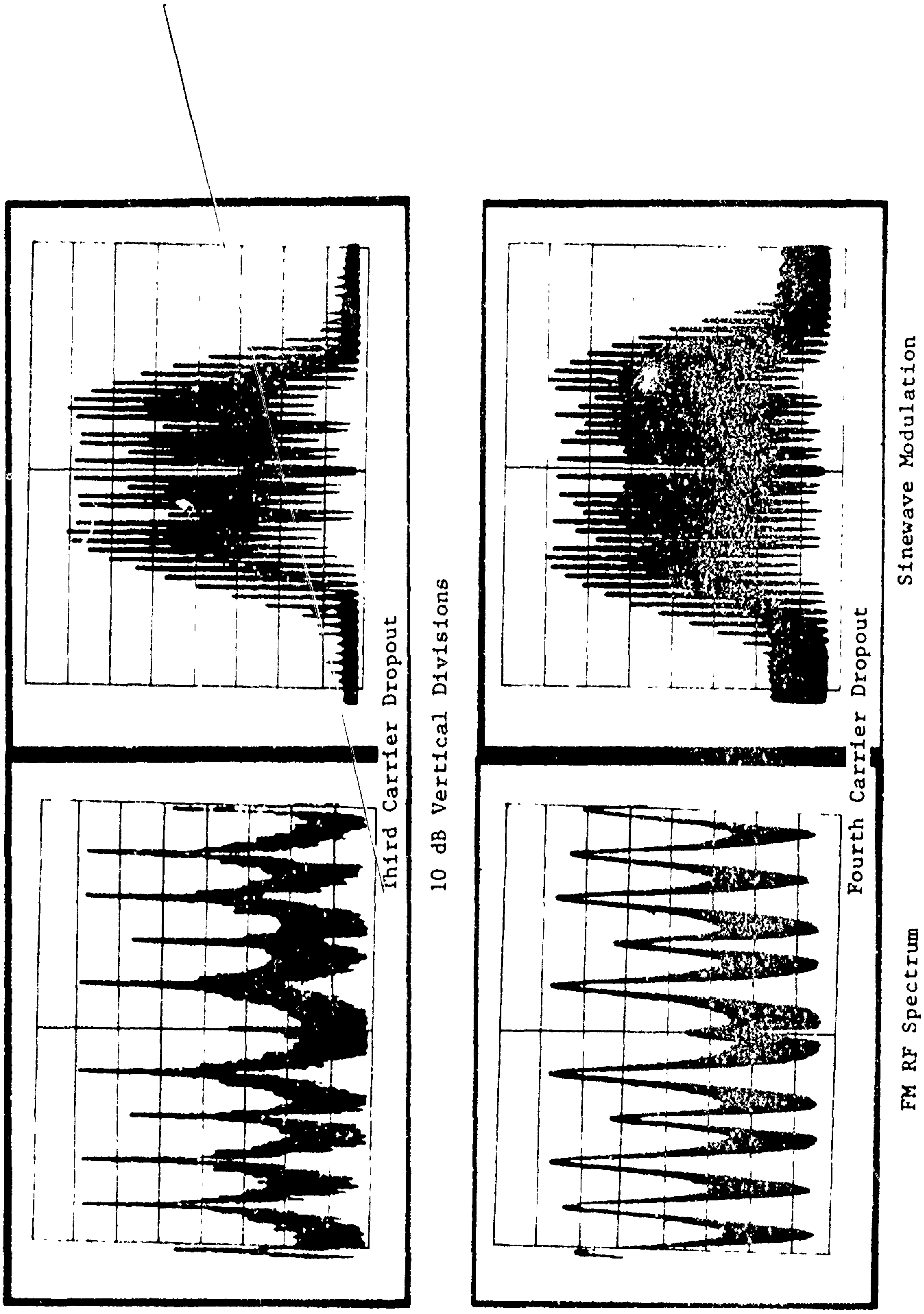


Figure 69
(cont.)

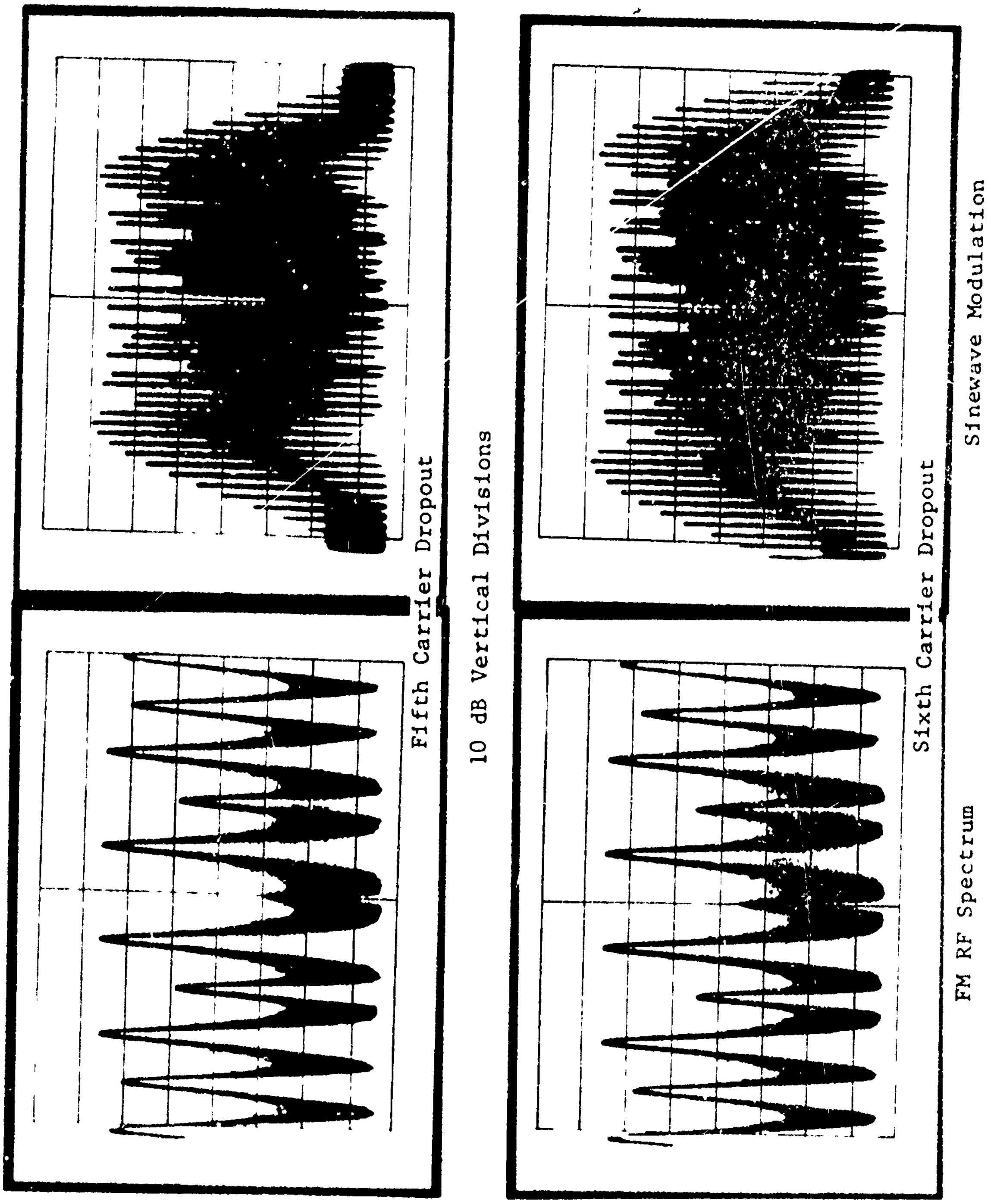


Figure 69
(cont.)

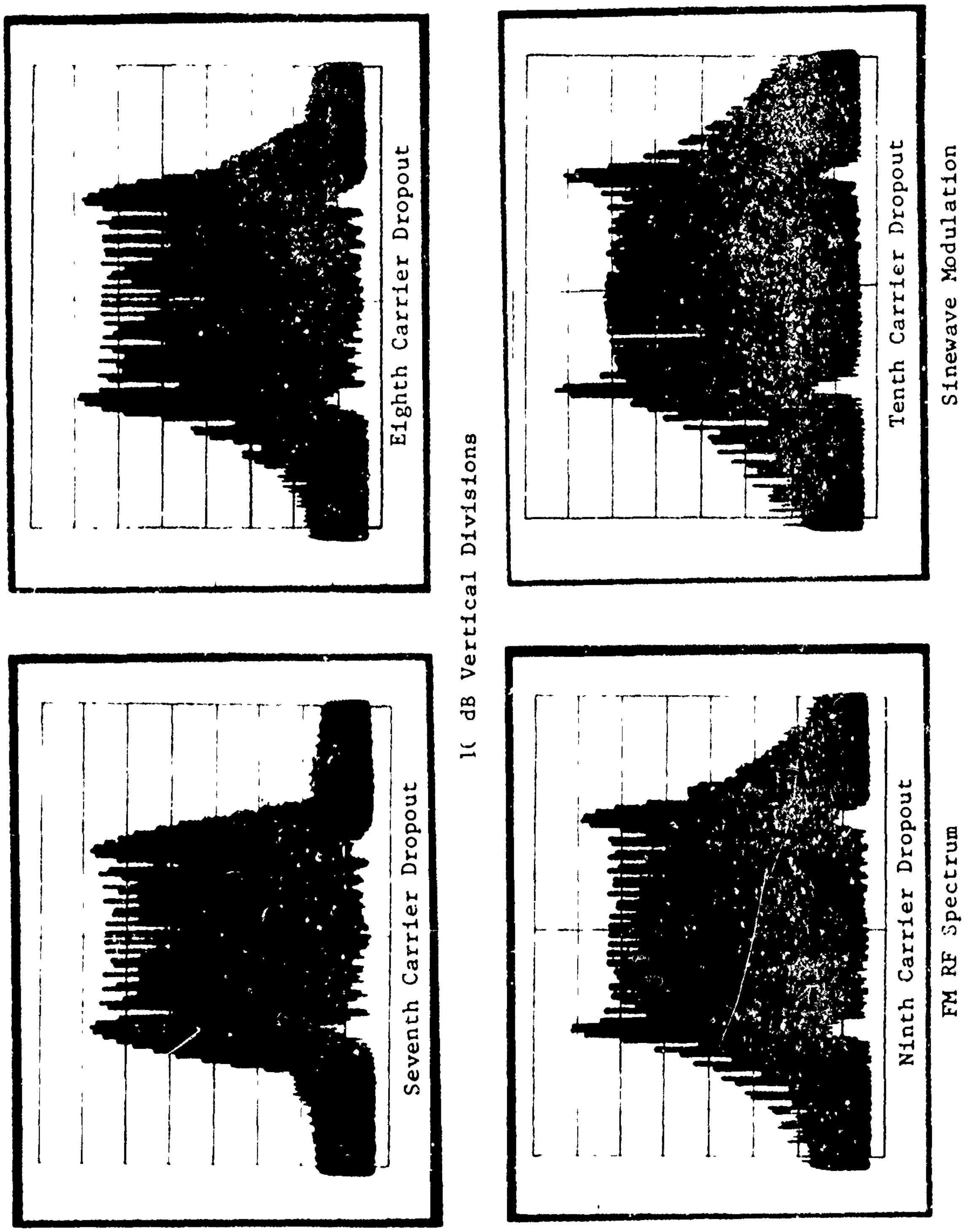


Figure 69
(cont.)

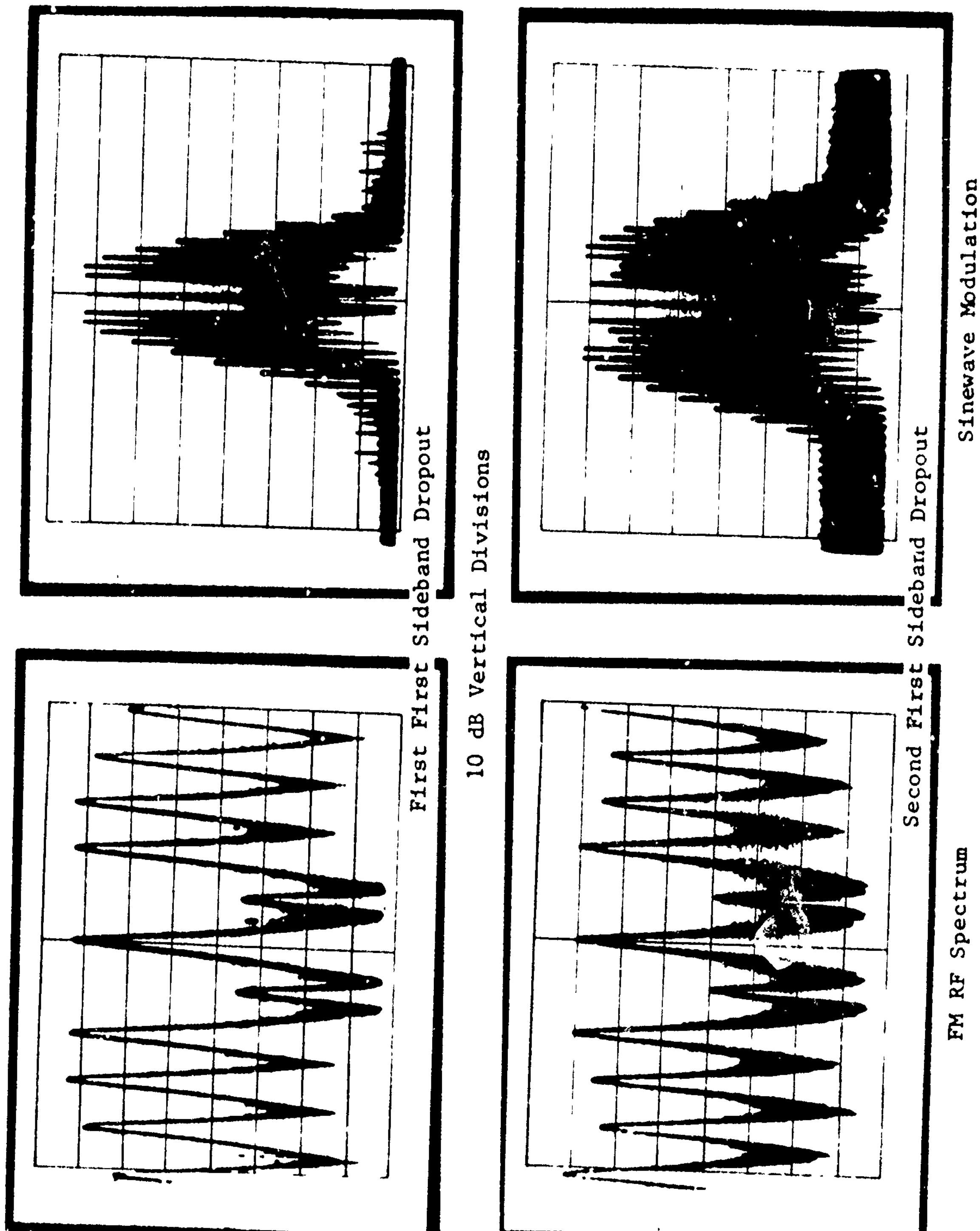


Figure 69
(cont.)

on just one side of the zero voltage axis or clipping the signal on both sides. The next page shows first carrier dropout using a sinewave with one side of the sinewave clipped to about one half its normal peak amplitude. The following page shows first sideband dropout for the same clipped sinewave. These pictures illustrate two features of unsymmertic clipping of sinewaves. Whenever the modulation circuitry becomes nonlinear in such a way as to cause the sinewave to become unsymmetric with regard to the zero voltage axis, the RF sideband spectrum will be unsymmetric with respect to the carrier frequency. Also, when sideband dropout is attempted, it will be observed that simultaneous dropout of both sidebands with the same modulation drive level will not be possible.

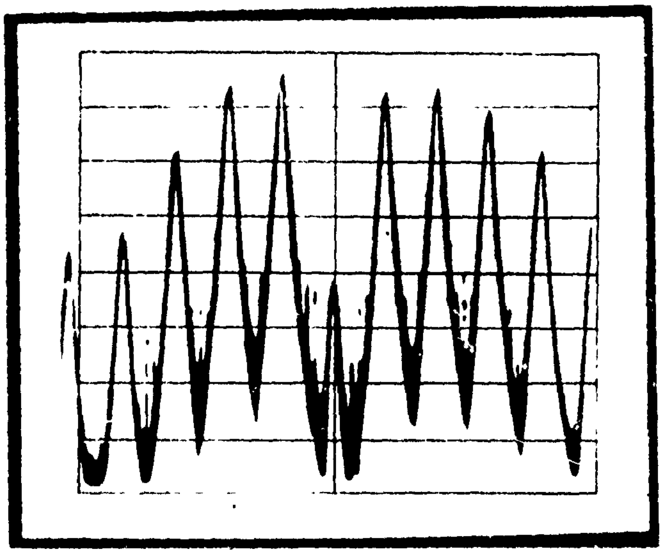
7.23 The next page shows the RF spectrum for first carrier dropout with a square wave. This situation would be approached with extreme symmetric clipping of the modulation circuitry. With square wave modulation, the carrier and sideband dropouts occur in much the same way as with sine wave modulation. However, the overall spectrum shape is considerably different. The next page shows low and high modulation index square wave modulation. Notice that the spectrum is consistently different than the spectrum for sine wave modulation.

7.24 There are other curious properties of carrier and sideband dropouts. For example, using sine waves, dropout first occurs for the carrier, then for first sideband, second sideband, third sideband, etc., in that order. As the modulator input drive level is increased, dropouts occur in the following order: first carrier (c), next, first sideband (1), then second sideband (2) and then second carrier dropout (c). In general, the order of dropouts is as follows:

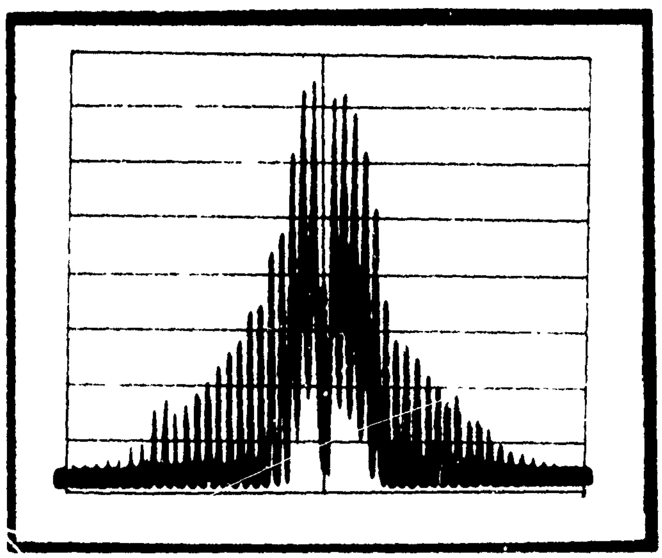
(c), (1), (2), (c), (3), (1), (4), (2), (c), (5), (3), (6), (1),
(4,7), (2), (c), (8), (5), (3), (1,9). (4), (9), (2), (7), etc.

7.25 For square wave modulation (severely clipped sine waves), however, dropout does not first occur at carrier frequency. First dropout occurs simultaneously for all odd order sidebands greater than one. Next, dropout occurs for the carrier and all even ordered sidebands greater than 2. In general, the dropouts occur in the following order:

(3, 5, 7, 9, 11, 13, 15, ...)
(c, 4, 6, 8, 10, 12, 14,...)
(1, 5, 7, 9, 11, 13, 15...)
(c, 2, 6, 8, 10, 12, 14,...)
(1, 3, 7, 9, 11, 13, 15,...)
(c, 2. 4, 8, 10, 12, 14,...)
(1, 3, 5, 9, 11, 13, 15,...)
(c, 2, 4, 6, 10, 12, 14,...)
(1, 3, 5, 7, 11, 13, 15,...)
(c, 2, 4, 6, 8, 12, 14,...) etc.



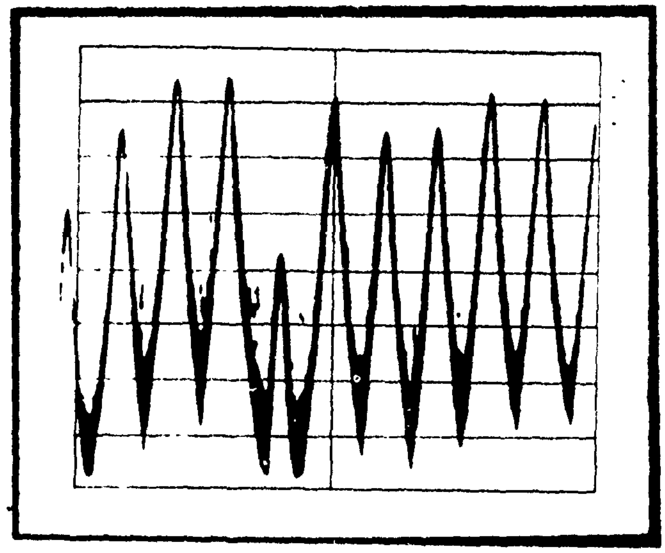
10 dB Vertical Divisions



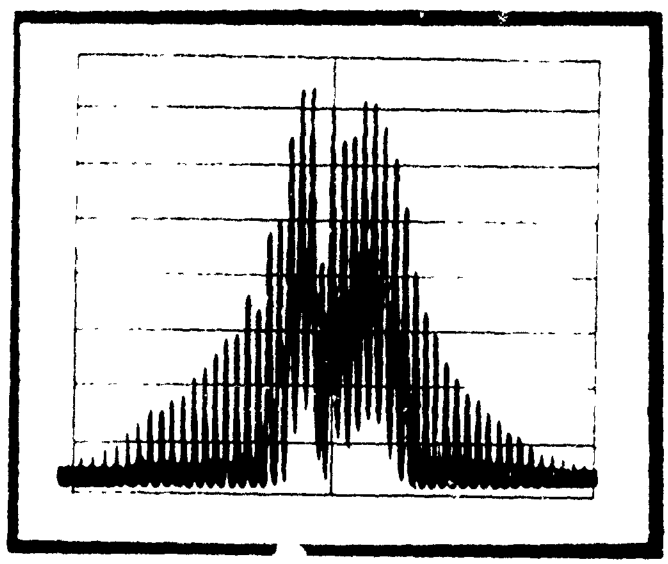
FM RF Spectrum

First Carrier Dropout
Modulation:
sinewave with one peak clipped

Figure 70



10 dB Vertical Divisions



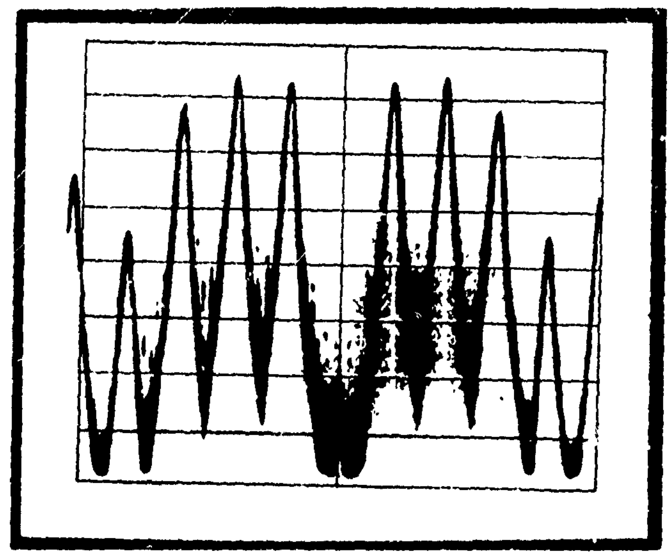
FM RF Spectrum

First First Sideband Dropout

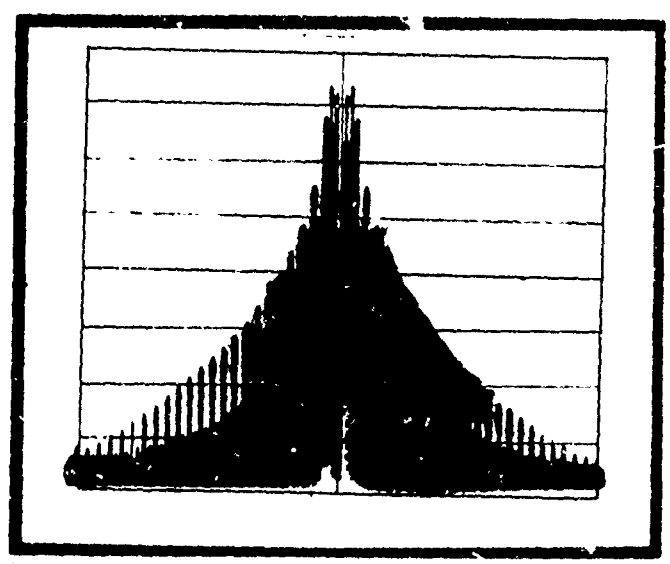
Modulation:

sinewave with one peak clipped

Figure 71

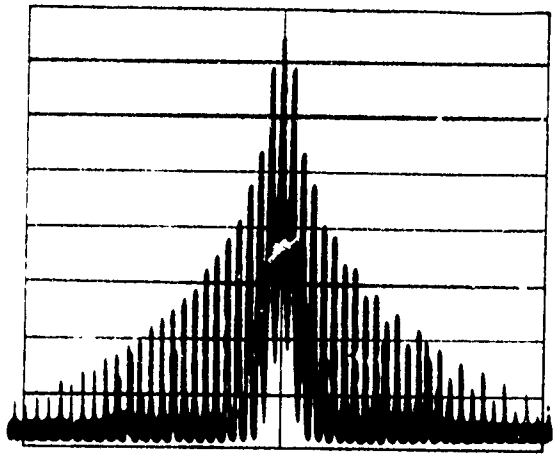


10 dB Vertical Divisions



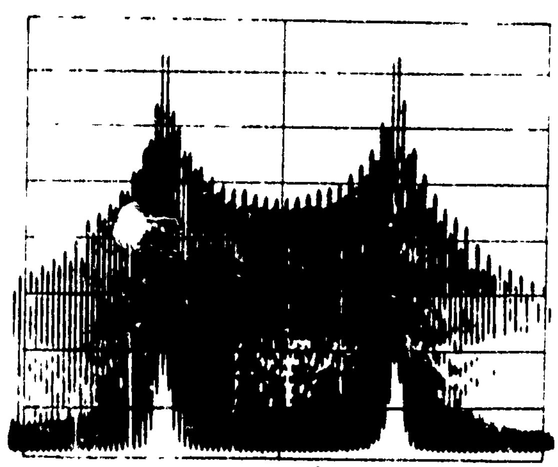
FM RF Spectrum
First Carrier Dropout
Square Wave Modulation

Figure 72



Low Modulation Index

10 dB Vertical Divisions



High Modulation Index

FM RF Spectrum

Square Wave Modulation

Figure 73

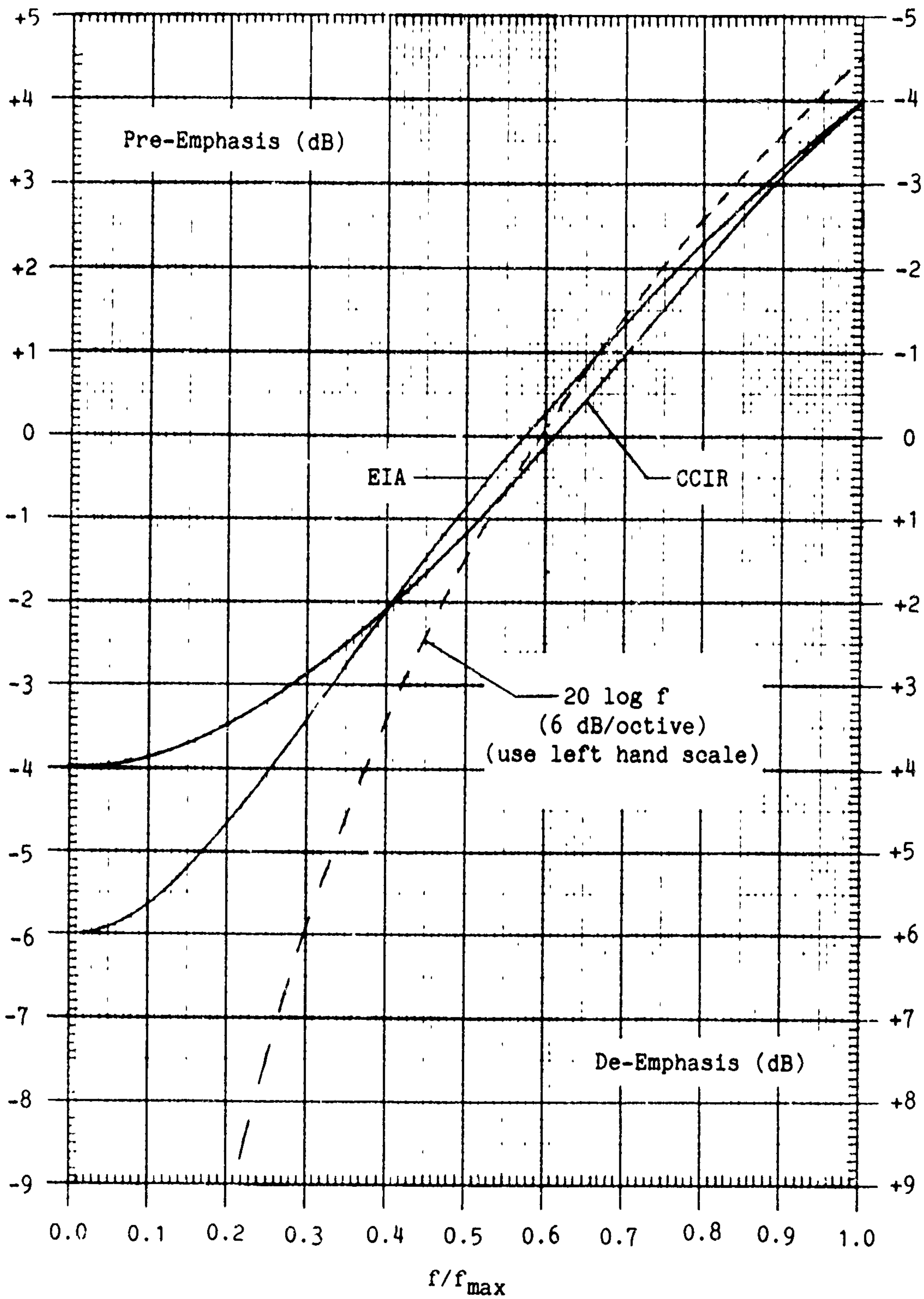
7.26 The above properties and other differences in the spectrum of sine and square wave modulated FM can be observed by studying the RF spectrum charts earlier in this report. These two simple cases illustrate the nonlinear character of FM modulation. Even for simple modulation waveforms, it is relatively complicated to determine what the modulated RF spectrum will be for an arbitrary modulator drive level.

Emphasis

7.27 As has been mentioned earlier, the slot noise out of the baseband of a radio in the linear portion (region C) of the quieting curve varies directly as frequency squared. This means that the higher the frequency of the 3.1 KHz FDMed telephone channel in the radio baseband, the more noise that channel will receive from the radio receiver. Specifically, for every doubling in baseband 3.1 KHz slot frequency (octave increase), the slot noise power (due only to thermal noise in the microwave receiver) increases by a factor of four. Hence, the thermal noise increases at a rate of 6 dB per octave (20 dB per decade) increase in baseband frequency.

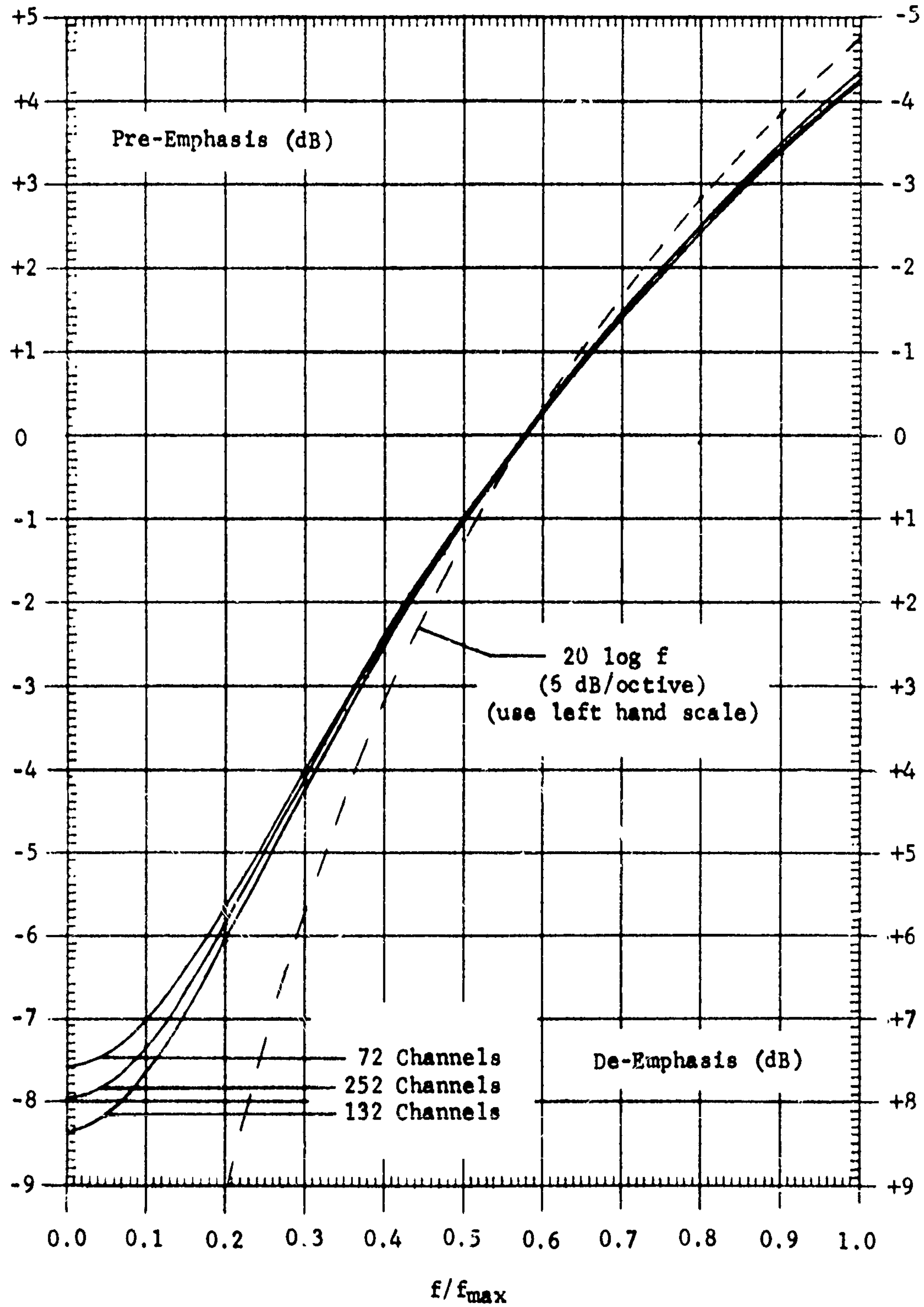
7.28 In an attempt to compensate for the undesirable increase in baseband slot noise as baseband frequency is increased, a rolloff network is often placed in the receiver output baseband circuitry. The network has a frequency response which goes down in frequency at approximately 6 dB per octave. Such networks are called de-emphasis networks. Of course, putting a de-emphasis network into the receiver circuitry radically affects the baseband frequency response of the receiver. To compensate for this, a pre-emphasis network is placed in the transmit baseband circuitry. To make sure that the overall baseband frequency response of the transmitter/receiver combination is uniform with changing baseband frequency, the emphasis networks are complimentary. This means that if de-emphasis frequency response is -1dB at a given frequency, the pre-emphasis value at the same frequency is +1dB. Therefore, if the dB frequency response of one of the networks is known, changing the sign of the dB value of the frequency response produces the frequency response of the other network. The frequency response of the two most common forms of emphasis (CCIR and EIA) are plotted on the next page. Also, plotted on the next page is the slot noise that will appear at the baseband of the radio if no de-emphasis is used. Notice that both of the emphasis networks do an excellent job of compensating for the thermal (6 dB per octave) noise over the top octave of the baseband.

7.29 Clearly the de-emphasis networks will effect the thermal noise performance of a microwave receiver. Baseband noise spectrums were run for typical microwave receivers with and without de-emphasis networks. Those spectrum plots follow on the next page. Notice that the baseband noise power for the LOS receiver is relatively flat with no RSL (C/N = $-\infty$dB). For a relatively high RSL (e.g., C/N = + 30 dB) the slot noise power is 6 dB per octave across the baseband. For the TROPO receiver, the frequency response of the CCIR de-emphasis network



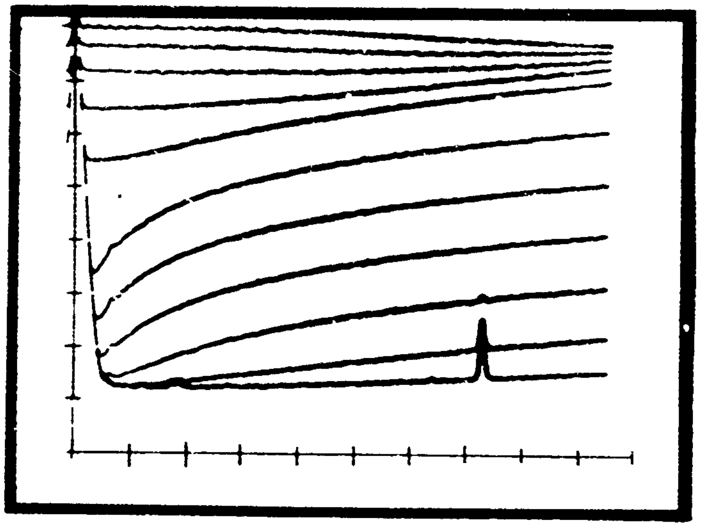
CCIR/EIA Emphasis Curves

Figure 74



REL Emphasis Curves

Figure 74 1



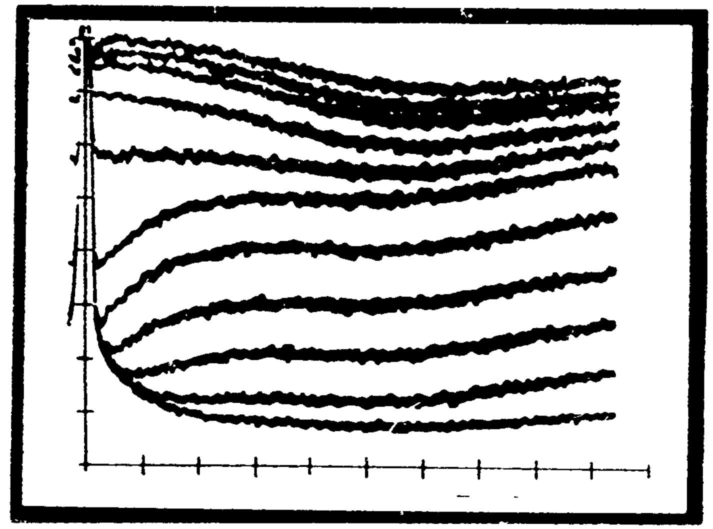
LOS M/W Receiver
no de-emphasis

Normal Baseband : 60 to 2660 kHz

Vertical : 10 dB/Div.
Horizontal: 0.5 MHz/Div.
(0 to 5 MHz)

Measurement Bandwidth = 30 kHz

C/N (dB) = -∞, 0, +3, +6, +10, +20, +30, +40, +50, +60, +70



TROPO M/W Receiver
CCIR de-emphasis

Normal Baseband : 12 to 252 kHz

Vertical : 10 dB/Div.
Horizontal · 50 kHz/Div.
(0 to 500 kHz)

Measurement Bandwidth = 3 kHz

C/N (dB) = -∞, 0, +3, +5, +8, +10, +20, +30, +40, +50, +60

Baseband Noise Spectrum for Various C/Ns

Figure 75

is clearly superimposed on the relatively flat baseband noise power when the receiver has no RSL. However, for relatively high RSLs, the baseband noise power is essentially flat over the top octave of the baseband.

7.30 To predict quieting curves and per channel rms deviation for systems with emphasis, it is necessary to determine the characteristics of emphasis networks. As EIA Standard RS-252-A (paragraph 2.10.1) and CCIR Documents of the XIII Plenary Assembly (Volume I, Recommendation 404-2, paragraph 2) point out, when pre-emphasis is used, the pre-emphasis characteristic should be such that the effective (total rms) deviation due to the composite multichannel baseband signal is the same with and without pre-emphasis. This means that the total power of the baseband signal will be the same out of pre-emphasis network as it was into the network (disregarding frequency independent circuit losses, of course). This is done so that the RF spectrum of a pre-emphasised transmitter will be approximately the same as that of an unpre-emphasised system. The pre-emphasisis network increases the level of a test tone at the high end of the baseband and decreases the level of the test tone at low baseband frequencies. At some frequency, however, the test tone level will pass through the pre-emphasis network unchanged (relative to its level if pre-emphasis had not been used). This frequency is called the pivot frequency, mean frequency, or crossover frequency of the pre-emphasis network. This report will call it pivot frequency. Given an arbitrary frequency response for a pre-emphasis characteristic, it is desirable to determine the pivot frequency of that network. Pre-emphasis relative to that frequency can then be determined. The pre-emphasis relative to pivot frequency is the pre-emphasis characteristic necessary for FM system noise calculations.

7.31 To determine the pivot frequency of a pre-emphasis network, it is necessary to determine the average power response of the network. If Pr(f) is the dB pre-emphasis characteristic (formula) for the network, let pr(f) be the power ratio response of that network.

$$Pr(f) = 10 \log pr(f)$$

$$pr(f) = \text{antilog}\ (Pr(f)/10)$$

7.32 The average power response of the network is just the integral of the power response of the pre-emphasis network integrated over the baseband frequency range of interest divided by the baseband bandwidth of interest. Formally, this is represented as:

$$pavg = \int_{fmin}^{fmax} pr(f)\, df \ / \ (fmax-fmin)$$

where pavg = average pre-emphasis power response
pr(f) = power response of the pre-emphasis network
fmin = lowest baseband frequency
fmax = highest baseband frequency

7.33 After determining the average power response, this value is equated to the pre-emphasis formula pr(f) and the equation solved for frequency. This frequency is pivot frequency.

7.34 The dB pre-emphasis response P(f) of the network relative to pivot frequency is:

$$P(f) = 10 \log (pr(f)/pavg)$$

7.35 If the signal power is equally spread over the frequency range of the baseband (uniform baseband signal power density) into the pre-emphasis network, the signal power out of the network will not be evenly spread over the baseband. However, the total power (signal power density integrated over the range of baseband frequencies), according to CCIR and EIA recommendations, will be the same at input and output of the pre-emphasis network (Actually, there will be some circuit loss associated with the pre-emphasis network which is not frequency selective. In practice, compensation for this will be made by lowering the Transmission Level Point (TLP) of the pre-emphasis output). The integral of pr(f)/pavg over the range of the baseband frequencies is unity. Therefore, it is the response of the pre-emphasis network which corresponds to the EIA and CCIR recommendations. The above method is essentially the same as one suggested by E. F. Reynolds in an unpublished report.

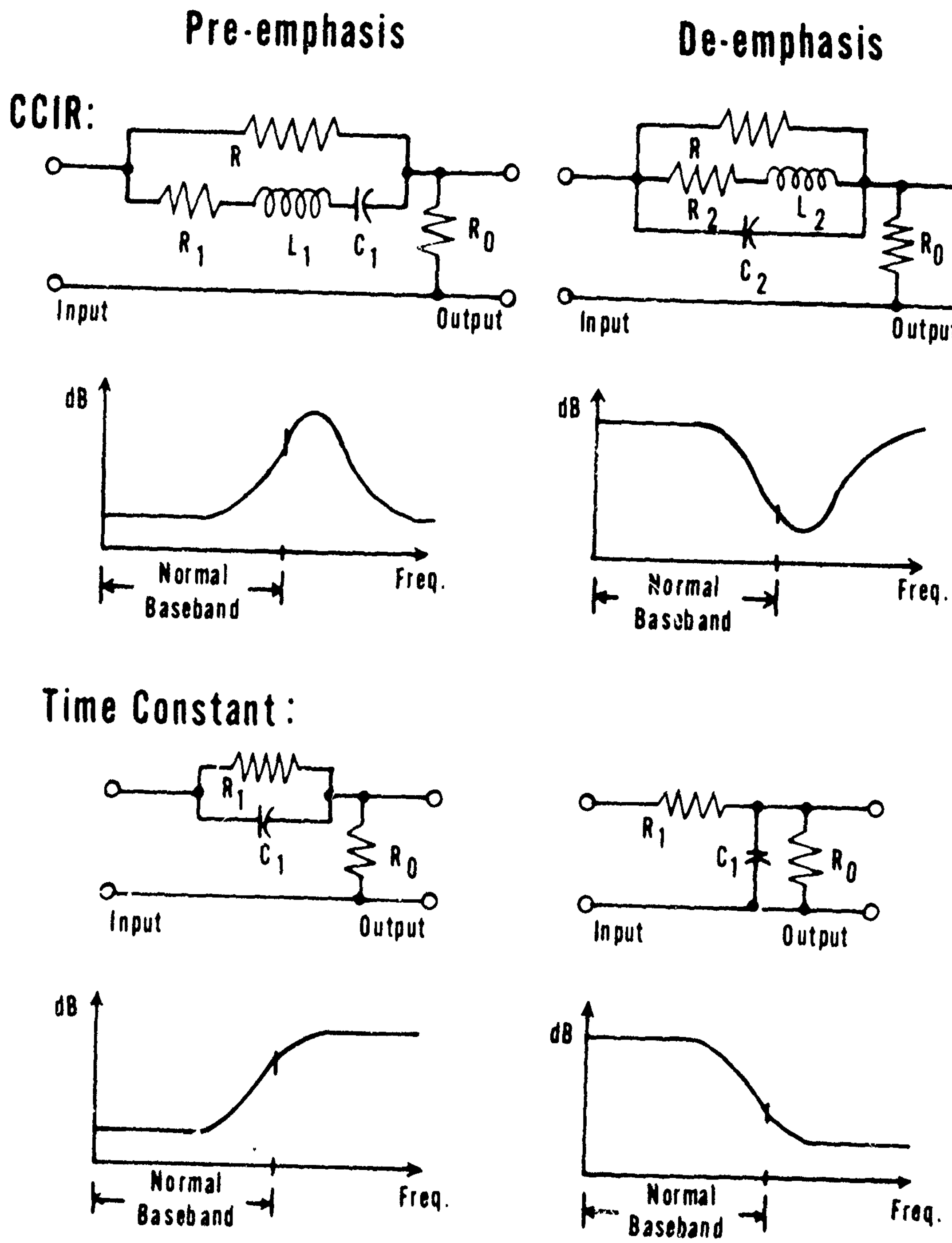
7.36 The two most common forms of emphasis are time constant and CCIR type. The circuit diagrams for these emphasis networks (if driven by low impedance source and driving a high impedance load) are given on the next page. The time-constant emphasis networks approximate integration and differentiation networks. The CCIR networks use parallel and series tuned circuits. For the CCIR networks, R is used to broaden (lower the Q) the response of the circuits and L and C are chosen so that the baseband frequencies of interest appear on the slope of the responses.

Time Constant Pre-emphasis:

7.37 Using the Laplacian operator S, the complex voltage transfer function of the pre-emphasis network is:

$$\frac{Vo}{VI} = \frac{Ro}{Z1 + Ro}$$

$$Z1 = R1 \parallel C1 = \frac{R1}{1 + s\, R1C1}$$



Emphasis Networks
(Circuit Diagrams and Power Responses)

Figure 76

For R much larger than Ro,

$$\frac{Vo}{V1} = \frac{Ro\ R1 - s^2(RoR1C1)^2}{R1^2 - s^2(RoR1C1)} + s\,\frac{RoR1^2C1}{R1^2 - s^2\ (RoR1C1)^2}$$

Letting $\tau = R_1C_1$ = time constant (seconds) where R1 has the units of megaohms and C is in microfarads,

$$\frac{V1}{V1} = \frac{Ro}{R1}\quad(1 + s\tau)$$

The power response of the pre-emphasis network is the product of the voltage transfer function times its complex conjugate.

$$\frac{Po}{PI} = \frac{Vo}{V1}\left(\frac{Vo}{V1}\right)^*$$

Letting S = j w to get the steady state response (Fourier transform) and neglecting the constant $(R_0/R_I)^2$,

$$\frac{Po}{P1} = 1 + (2\pi)^2\tau^2\ f^2 = 1 + 39.5\tau^2f^2$$

Where

τ = time constant (in seconds)
f = baseband frequency (in Hertz)

$$pavg = \frac{\int_{fmin}^{fmax} (1 + (2\pi\tau)^2\ f^2)\ df}{fmax - fmin}$$

$$= 1 + \frac{(2\pi\tau)^2}{3}\ (fmax^2 + fmax\ fmin + fmin^2)$$

This can be approximated by

$$pavg = 1 + \frac{(2\pi\tau)^2}{3}\ fmax^2$$

The error of this approximation is a function of baseband bandwidth. This error (in dB) will be shown in the Pre/De-Emphasis Approximation Error Analysis table which follows.

Equating pavg to pr(f)

$$1 + (2\pi\tau)^2 f = 1 + \frac{(2\pi\tau)^2}{3}\, fmax^2$$

yields pivot frequency

$$f_p = \frac{1}{\sqrt{3}}\, fmax = 0.577\ fmax$$

1.38 Since pavg is an approximation which has baseband bandwidth dependent error, this error is transfered to the above approximations of pivot frequency. The error (as %) associated with this approximation is also listed in the Pre/De-Emphasis Approximation Error Analysis Chart which follows.

The dB pre-emphasis network frequency response becomes

$$P(f) = 10 \log\ Pr(f)/pavg$$

$$= 10 \log \frac{1 + (2\pi\tau)^2 f^2}{1 + \frac{(2\pi\tau)^2}{3} fmax^2}$$

$$= 10 \log \frac{1 + 39.48\ {}^2 f^2}{1 + 13.16\,\tau^2 fmax^2}$$

where

τ = time constant (microseconds)
f = baseband frequency (megahertz)
fmax = highest frequency in baseband (megahertz)

since specifying f in megahertz and τ in microseconds yields the same result as specifying f in Hertz and τ in seconds. This is desirable since τ is usually specified in microseconds.

EIA pre-emphasis:

The EIA recommends a particular form of time constant pre-emphasis. The following values are recommended:

$$\tau = R_1 C_1 = 3/(2\pi\, fmax)$$

$$R_1 > 9R_0$$

Taking advantage of the previous result,

$$fp = 0.577 \, fmax$$

$$P(f) = 10 \log \frac{1 + 9(f/fmax)^2}{4}$$

$$= 10 \log (0.250 + 2.25 \, (f/fmax)^2)$$

7.39 The errors associated with these approximations are listed in the following error analysis chart.

7.40 The EIA recommends the use of the following formulas for their pre-emphasis network

$$fp = 0.577 \, fmax$$

$$P(f) = 10 \log (0.251 + 2.26 \, (f/fmax)^2)$$

7.41 No mention is made by EIA of the error associated with these formulas. However, the EIA standard does specify that the frequency response of the actual pre-emphasis networks shall not depart from the preceeding formula by more than $\mp$ (0.1 + 0.05 (f/fmax)) dB.

CCIR pre-emphasis:

The recommended formula for CCIR pre-emphasis is

$$Pr(f) = 5 - 10 \log \left[1 + \left(\frac{6.90}{1 + \left(\dfrac{5.25}{\left[\dfrac{1.25 \, fmax}{f} - \dfrac{f}{1.25 \, fmax} \right]^2} \right)} \right) \right]$$

7.42 CCIR recommends that actual pre-emphasis networks approximate the preceeding frequency response with error no greater than $\mp$ (0.1 + 0.05 (f/fmax)) dB error. The preceeding formula is rather cumbersome. Direct algebraic manipulation results in the following equivalent form.

$$Pr(f) = 10 \log \frac{0.400 + 0.832(f/fmax)^2 + 0.164(f/fmax)^4}{1.000 - 0.855(f/fmax)^2 + 0.410(f/fmax)^4}$$

7.43 This formula is certainly more convenient than the original formula. However, it would be convenient if a simpler form could be found. Also, since the factor inside the brackets must be integrated to determine pavg, a single polynomial would be most desirable.

EIA recommends the following formula for CCIR pre-emphasis:

$$Pr = 10 \log (0.401 + 1.34 (f/fmax)^2 + 0.77 (f/fmax)^4)$$

7.44 This formula differs from the CCIR recommendation by as much as 0.14 dB. This is an unacceptable approximation since its error is nearly the worst allowable for an actual pre-emphasis network.

7.45 The most direct approach to producing a polynomial would be to divide the demoninator of the modified CCIR formula into the numerator. This produces the following:

$$Pr(f) = 10 \log (0.40000 + 1.17387 (f/fmax)^2 + 1.00328 (f/fmax)^4 + 0.37666 (f/fmax)^6 - 0.08902 (f/fmax)^8 - 0.23036 (f/fmax)^{10} - 0.16042 (f/fmax)^{12} - \ldots)$$

7.46 The preceeding infinite series is inefficient. Seven terms are necessary to reduce the maximum error to 0.07dB. Use of fewer terms rapidly leads to excessive error.

7.47 To yield a more efficient formula, the CCIR curve was approximated using least squared error curve fitting techniques. To take advantage of the smooth nature of the CCIR curve and minimize the number of terms, before performing a least squares fit, the data was predistorted so as to produce only even ordered multiples of the term (f/fmax). Odd ordered terms would have been useless since they produce kinks which the CCIR curve obviously does not have.

7.48 The following approximations and their maximum error over the range f/fmax = 0 to f/fmax = 1 are as follows:

$$Pr(f) = 10 \log (0.388 + 1.34 (f/fmax)^2 + 0.853 (f/fmax)^4)$$

maximum error = 0.13 dB

$$Pr(f) = 10 \log (0.406 + (f/fmax)^2 + 1.82 (f/fmax)^4 - 0.698 (f/fmax)^6)$$

maximum error = 0.07 dB

$$Pr(f) = 10 \log (0.400 + 1.20 (f/fmax)^2 + 0.801 (f/fmax)^4 + 1.03 (f/fmax)^6 - 0.913 (f/fmax)^8)$$

maximum error = 0.01 dB

To determine pavg, the five term approximation was used.

$$\int_{fmin}^{fmax} pr(f)\, df = \int_{fmin}^{fmax} (0.3995 +$$
$$1.196\,(f/fmax)^2 + 0.8009\,(f/fmax)^4 +$$
$$1.032\,(f/fmax)^6 - 0.9128\,(f/fmax)^8\,)\, df$$
$$= 0.3995\,(fmax-fmin) + 0.3987\,(fmax - fmin\,(fmin/fmax)^2)$$
$$+ 0.1608\,(fmax - fmin\,(fmin/fmax)^4)$$
$$+ 0.14743\,(fmax - fmin\,(fmin/fmax)^6\,)$$
$$- 0.10142\,(fmax - fmin\,(fmin/fmax\,)^8\,)$$

This answer can be approximated (with maximum error of 0.02 db) by the following:

$$\int_{fmin}^{fmax} pr(f)\, df = 1.004\, fmax - 0.3995\, fmin$$

Therefore

$$pavg = \frac{\int_{fmin}^{fmax} pr(f)\, df}{fmax - fmin} = \frac{fmax - 0.4\, fmin}{fmax - fmin}$$

$$= \frac{1 - 0.4\,(fmin/fmax)}{1 - (fmin/fmax)}$$

Obviously for a sufficiently wide baseband,

$$pavg = 1.0$$

7.49 The above assumption will be used in the following results. The pre-emphasis error (in dB) and the pivot frequency error (in %) associated with that assumption have been summarized in the following table. Setting pavg equal to pr(f) and solving for f yields the following pivot frequency:

$$fp = 0.613\, fmax$$

CCIR recommends

$$fp = 0.608\, fmax$$

7.50 If pavg is considered equal to one, pr(f) is equal to p(f). Therefore, the preceeding formulas for pr(f) are the formulas for pre-emphasis relative to pivot frequency.

7.51 The following page lists pivot frequencies for CCIR, EIA, and time constant emphasis networks. The 0.608 and 0.577 factors were used to generate the chart.

7.52 To determine the error associated with the previous approximations, error analysis was performed for the three most common types of pre-emphasis. The various columns of the following error analysis chart were derived using the following formulas.

$$\% \text{ errors} = \frac{\text{actual pivot frequency} - \text{approximate pivot frequency}}{\text{actual pivot frequency}}$$

dB error = actual pre-emphasis relative to pivot frequency (dB)
- approximation of pre-emphasis relative to pivot frequency (dB)

7.53 Actual analysis of the error associated with time constant pre-emphasis requires knowledge of the time constant. For purposes of the error analysis chart, a time constant of 10 microseconds was arbitrary used. Therefore, the time constant column should be used only to observe the rough trend of the errors involved.

REL Pre-emphasis:

7.54 Given a phase modulator, it is desired to determine an equivalent pre-emphasis network to allow comparison of the transmitter as a system with more conventional FM modulators. To show how this can be done, the following example is given:

7.55 The serrasoid modulator produces an output by changing the relative position (in time) of a trapezoidal waveform based on the instantaneous voltage of the baseband signal applied to the modulator. Before transmission, the trapezoidal waveforms are smoothed (filtered) into sinewaves. The position in time (i.e., phase) of the sinewave is proportional to the position in time of the trapezoid. Since the position of the trapezoid is proportional to input voltage, the phase of the final sinewave is proportional to the input voltage; therefore, the phase of the final sinewave is proportional to the input voltage. By definition of phase modulation, the serrasoid must be a phase modulator (Also, para 4-96, page 4-17, of TO 31R5-2FRC39-252 specifically states that the serrasoid is a phase modulator.).

7.56 It was previously mentioned that a frequency modulator preceeded by a differentiation network produces exactly the same modulated RF output signal as a phase modulator. Instead of pre-emphasis networks, serrasoids are preceded by corrector networks. Therefore, all that is necessary to obtain the equivalent pre-emphasis network is to determine the frequency (voltage squared or "power") response of the corrector network and multiply that by the power response of a differentiation network. The equivalent pre-emphasis network is necessary to predict

VF Channels (number)	Baseband (kHz)	Pivot Frequency (kHz) CCIR	EIA (Time Constant)
12	12-60	36.5	34.6
24	12-108	65.7	62.3
36	12-156	94.8	90.0
48	12-204	124	118
48	60-252	153	145
60	12-252	153	145
60	60-300	182	173
72	12-300	182	173
120	60-552	336	319
132	12-552	336	319
240	60-1052	640	607
252	12-1052	640	607
300	60-1300	790	750
420	60-1796	1092	1036
600	60-2660	1617	1535
900	316-4188	2546	2416
960	60-4028	2449	2324
1200	316-5564	3383	3210
1260	60-5564	3383	3210
1800	316-8204	4988	4734
2400	316-11404	6934	6580

Pivot Frequencies

Table 30

VF Channels (number)	Baseband (kHz)	Time Constant %	Time Constant dB	EIA %	EIA dB	CCIR %	CCIR dB
12	12-60	19.3	0.93	15.3	0.72	13.2	0.61
24	12-108	11.0	0.51	8.5	0.38	7.4	0.33
36	12-156	7.6	0.35	5.8	0.26	5.2	0.23
48	12-204	5.9	0.26	4.5	0.20	4.0	0.18
	60-252	22.8	1.12	18.1	0.87	15.7	0.74
60	12-252	4.8	0.21	3.6	0.16	3.3	0.15
	60-300	19.4	0.93	15.3	0.72	13.2	0.61
72	12-300	4.0	0.18	3.0	0.13	2.9	0.13
120	60-552	10.8	0.49	8.3	0.38	7.2	0.32
132	12-552	2.17	0.10	1.6	0.07	1.7	0.08
240	60-1052	5.69	0.25	4.3	0.19	3.9	0.17
300	60-1300	4.61	0.20	3.5	0.15	3.2	0.14
420	60-1796	3.3	0.15	2.5	0.11	2.5	0.11
600	60-2660	2.3	0.10	1.7	0.07	1.8	0.08
900	316-4188	7.5	0.34	5.7	0.26	5.1	0.23
960	60-4028	1.5	0.07	1.1	0.05	1.3	0.06
1200	316-5564	5.7	0.25	4.3	0.19	3.9	0.17
1260	60-5564	1.1	0.05	0.8	0.04	1.1	0.05
1800	316-8204	3.8	0.17	2.9	0.13	2.8	0.12
2400	316-11404	2.8	0.12	2.1	0.09	2.1	0.09

Pre/De-Emphasis Approximation Error Analysis

Table 31

quieting curves for the companion receivers which will have complementary de-emphasis networks. The circuit diagram of the HF baseband circuitry (including the corrector network) of the serrasoid modulator is shown on the next page. The serrasoid has LF and HF baseband sections. Analysis of the LF circuitry is similar to the HF circuitry analysis. Assuming V5 interelectrode capacities have been sufficiently isolated by R46-48 and R45+44 so as to be insignificant, the baseband circuitry can be reduced (using conventional circuit analysis) to the simplified circuit diagram on the following page.

$$ZC16 = 1/(jwC16) = j\,\frac{1}{2\pi fC16}$$

$$= -j\,\frac{1}{2\pi f_{Hz}\,(300X10^{12})} = -j\,\frac{10^{10}}{6\pi f} = -j\,\frac{5.3052X10^{8}}{f_{Hz}}$$

$$R+ZC16 = R-j\,\frac{5.3052X10^{8}}{f}$$, R in ohms, f in Hertz

$$\frac{Vout}{Vin} = \frac{R+ZC16}{R32+R+ZC16}$$

$$\frac{Pout}{Pin} = \frac{Vout}{Vin}\;\frac{Vout^{*}}{Vin}$$

*Complex Conjugate

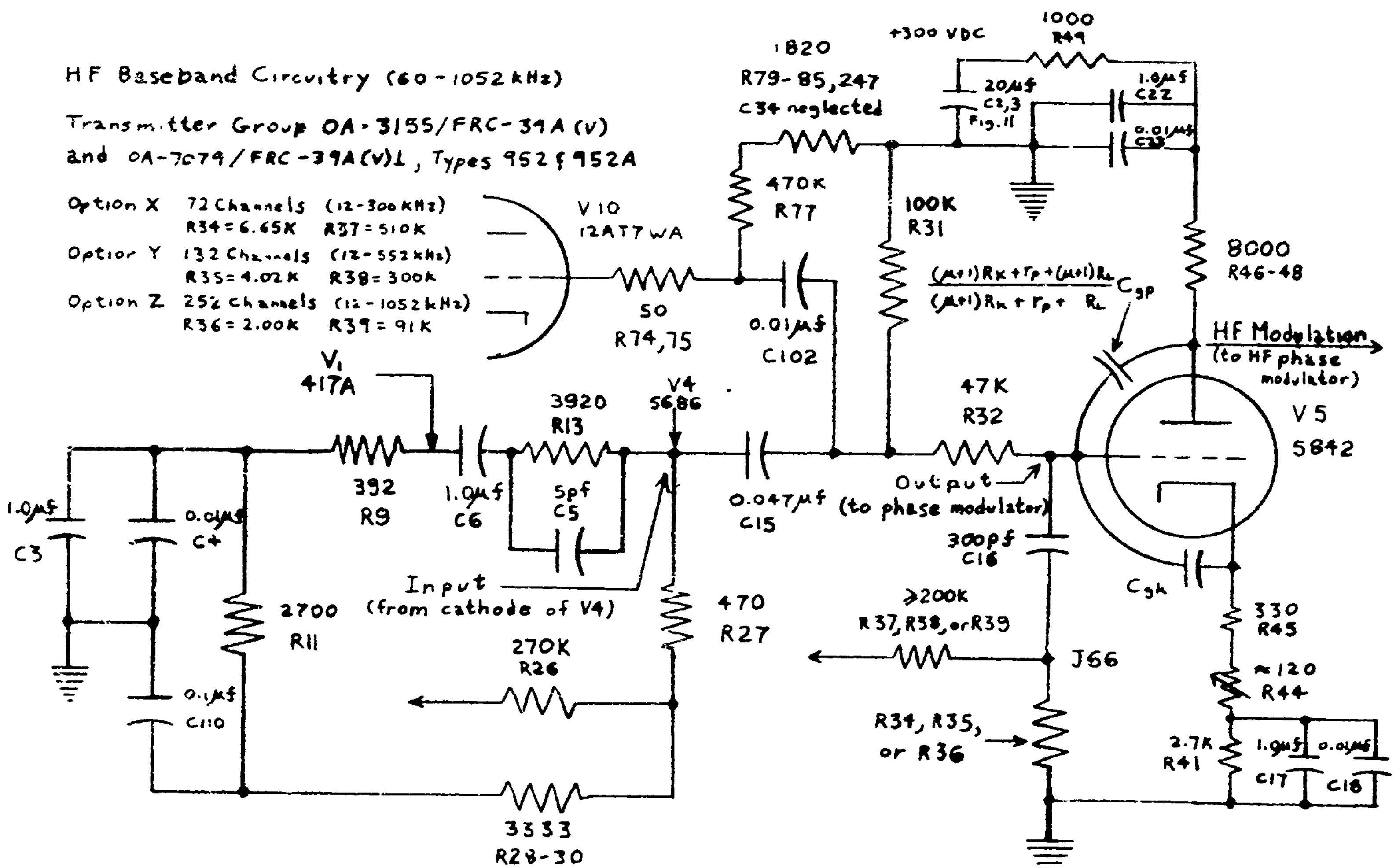
$$= \frac{2.8145X10^{17}}{(4.7X10^{4}+R)^{2}}\left[\frac{1}{f^{2}} + \frac{R^{2}}{2.8145X10^{17}}\right]$$

using the approximation

$$1+\frac{2.8145X10^{17}}{(4.7x10^{4}+R)^{2}f^{2}} \approx 1$$

7.57 Under worst case conditions (f=60KHz, R=0), the preceding approximation leads to a maximum error of 0.15 dB in the final result. The error introduced by this approximation is less than 0.05 dB for baseband frequencies greater than 105 kHz.

7.56 Multiplying the previous response by Kf^2, the power response of a differentiation network (K is just a constant to compensate for the gain between the corrector network and the modulator), yields the following:

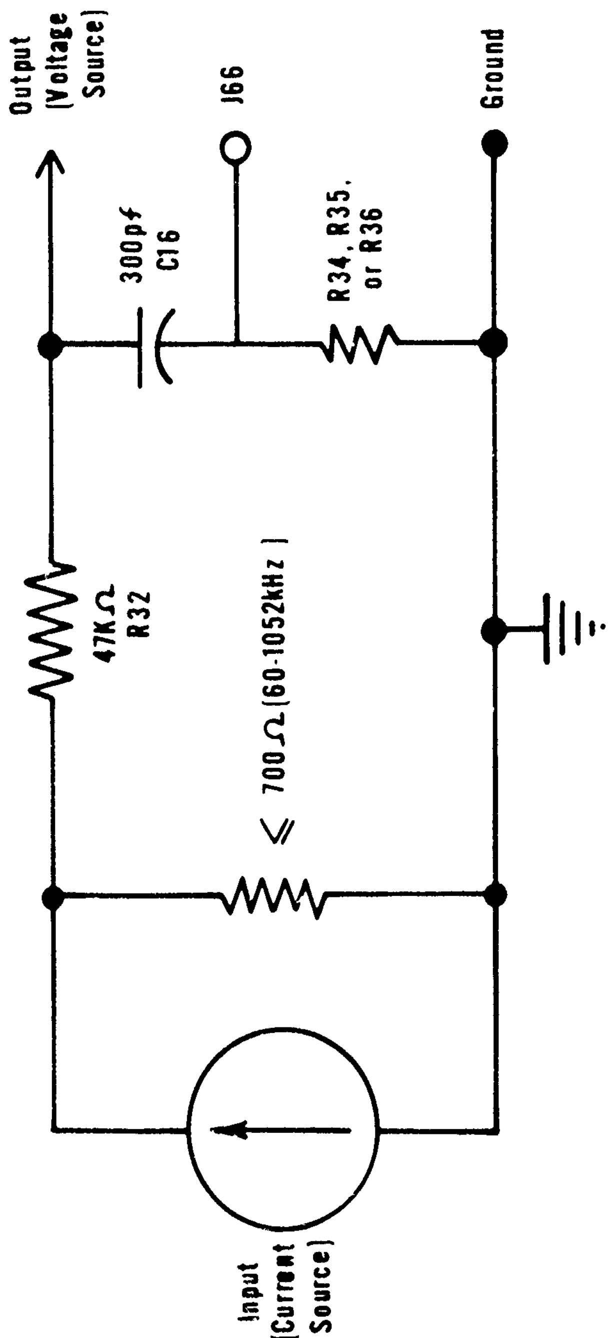


$$\frac{(\mu+1)R_k + r_p + (\mu+1)R_L}{(\mu+1)R_k + r_p + R_L} C_{gp}$$

Serrasoid Type Phase Shift Modulator

Input Baseband Circiutry

Figure 77



Serrasoid Type Phase Shift Modulator
Simplified Input Baseband Circuitry

Figure 78

$$\frac{Pout}{Pin} = \frac{2.8145\times10^{17}}{(4.7\times10^{4}+R)^2} K \left[1+ \frac{R^2 f^2}{2.8145\times10^{17}} \right]$$

7.59 Disregarding the constants in the equation (This is appropriate since later we will be taking the ratio of this equation with itself for two different frequencies. If the constants were retained, they would cancel at that point. Disregarding the constants now is merely a convenience.), the equation becomes the following:

$$\frac{Pout}{Pin} = 1 + \frac{R^2}{2.8145\times10^{17}} f^2$$

This is quite similar to the time constant pre-emphasis equation. Equating the two,

$$\frac{Pout}{Pin} = 1 + \frac{R^2}{2.8145\times10^{17}} f^2 = 1 + 39.478\,\tau^2 f^2$$

where τ in seconds, f in Hertz, and R in ohms

$$\tau = \frac{R^2}{2.8145\times10^{17}\,(39.478)} = \frac{R}{3.3333\times10^{9}} = \frac{(R/10^3)\times10^6}{3.3333}$$

If we redefine R in kilohms and τ in microseconds, the equation for τ becomes

$$\tau = \frac{R(k\Omega)}{3.333(\mu sec)}$$

Knowing that the equivalent pre-emphasis is time constant type allows us to write

$$P(f) = 10 \log \frac{1+39.478\,\tau^2 f^2}{1+13.159\,\tau^2 f_{max}^2}$$

Where $\tau = \frac{R}{3.3333}$

and

P(f) = pre-emphasis (in dB) relative to pivot frequency

τ = time constant (in μsec)

f = baseband frequency of interest (in MHz)

fmax = highest baseband frequency (in MHz)

R = resistance of R34, R35, or R36 (in KΩ).

A convenient alternate form of the above formula is the following:

$$P(f) = 10 \log \frac{1+39.478\tau^2 fmax^2 \ (f/fmax)^2}{1 + 13.159\tau^2 fmax^2}$$

Consider the various capacity systems:

72 channel system
(12 to 300 KHz)

fmax = 0.300 MHz = 300 kHz
R = R34 = 6.65 KΩ

τ = 6.65/3.3333 = 2.00 (μsec)

$$P(f) = 10 \log \frac{1 + 39.478(2.00)^2(0.300)^2 \ (f/fmax)^2}{1 + 13.159 \ (2.00)^2 \ (0.300)^2}$$

$$= 10 \log \frac{1 + 14.212 \ (f/fmax)^2}{5.737}$$

$$= 10 \log (0.17430 + 2.4772 \ (f/fmax)^2) \ (dB)$$

132 channel system
(12-552 KHz)

fmax = 0.552 MHz = 552 kHz
R = R35 = 4.02 KΩ

τ = 4.02/3.3333 = 1.21 (μsec)

$$P(f) = 10 \log \frac{1 + 39.478 \ (1.21)^2(0.552)^2(f/fmax)^2}{1 + 13.159 \ (1.21)^2 \ (0.552)^2}$$

$$= 10 \log \frac{(1 + 17.612 \ (f/fmax)^2 \ )}{6.870}$$

$$= 10 \log (0.14555 + 2.5634 \ (f/fmax) \ ) \ \ (dB)$$

252 channel system (12-1052 KHz)

fmax = 1.052 MHz
R=R36 = 2.00 KΩ

τ =2.00/3.333 = 0.600(μsec)

$$P(f) = 10 \log \frac{1 + 39.478(0.600)^2(1.052)^2(f/fmax)^2}{1 + 13.159 \ (0.600)^2 \ (1.052)^2}$$

$$= 10 \log \frac{1+15.729\ (f/fmax)^2}{6.2427}$$

$$= 10 \log (0.16019 + 2.5196\ (f/fmax)^2\ )$$

7.60 One last bit of information is needed. To predict quieting curve performance, it is necessary to know the pre-emphasis relative to pivot frequency. If per channel rms deviation is measured with pre-emphasis strapped out (in this case, J66 strapped to ground), a correction factor will be needed to determine the deviation value to be used in the quieting curve prediction. That factor is the ratio of the circuit power gain (pre-emphasis boost) at pivot frequency with pre-emphasis in over the circuit power gain at pivot frequency with pre-emphasis strapped out. For all modern pre-emphasis networks (CCIR, EIA), this power ratio is one (0 dB). For the older time constrant pre-emphasis networks (including the equivalent serrasoid pre-emphasis network), the ratio is considerably different than one.

7.61 Shorting J66 to ground is the same as forcing R to be zero. If R is zero, τ is zero. Therefore, the previously mentioned ratio Pp(f) expressed in dB is the following:

$$Pp(f) = 10 \log \frac{1+39.478\tau^2 fp^2}{1+39.478(0)fp^2}$$

$$= 10 \log (1+39.478\tau^2 fp^2)$$

with

fp = pivot frequency = 0.577 fmax

The Pp(f) factors for the various capacity radios follows:

72 channel capacity radio:

$$Pp(f) = 10 \log (1+14.212\ (0.577\ fmax/fmax)^2)$$

$$= 10 \log(5.7315) = +7.58\ dB$$

132 channel capacity radio:

$$Pp(f) = 10 \log (1+17.612\ (0.577\ fmax/fmax)^2\ )$$

$$= 10 \log (6.8635) = +\ 8.37\ dB$$

252 channel capacity radio:

$$Pp(f) = 10 \log (1+\ 15.729\ (0.577\ fmax/fmax)^2\ )$$

$$= 10 \log (6.2366) = +7.95\ dB$$

IF Bandwidth

7.62 There are two questions to consider with regard to IF bandwidth. The first is "How does the shape of the IF frequency response effect noise prediction and measurement?" The second is "How can IF bandwidth be estimated based on equipment parameters?"

7.63 First, consider the IF bandwidth measurement problem. In the theoretical developments of FM radio noise performance, it has always been assumed that the noise bandwidth was being used (see noise figure section for definition). From a practical point of view, however, the 3 dB (half power) bandwidth is usually measured. Does taking the 3 dB bandwidth as equal to the noise bandwidth cause significant error in our calculations?

7.64 To simplify matters, we will consider the two worst cases. It is the author's opinion that rectangular and Gaussian IF responses are, respectively, the sharpest and the broadest IF responses that will be encountered in practice.

7.65 A rectangular IF would have the following response (neglecting any gain or loss in the IF).

$$H(f) \begin{cases} 1, & \text{for any frequency within the bandpass of the IF} \\ 0, & \text{for any frequency outside the bandpass of the IF} \end{cases}$$

7.66 Clearly, the noise bandwidth and 3dB bandwidth are identical (in a limiting sense).

7.67 A Gaussian IF response would have the following power response (again neglecting any gain or loss in the IF):

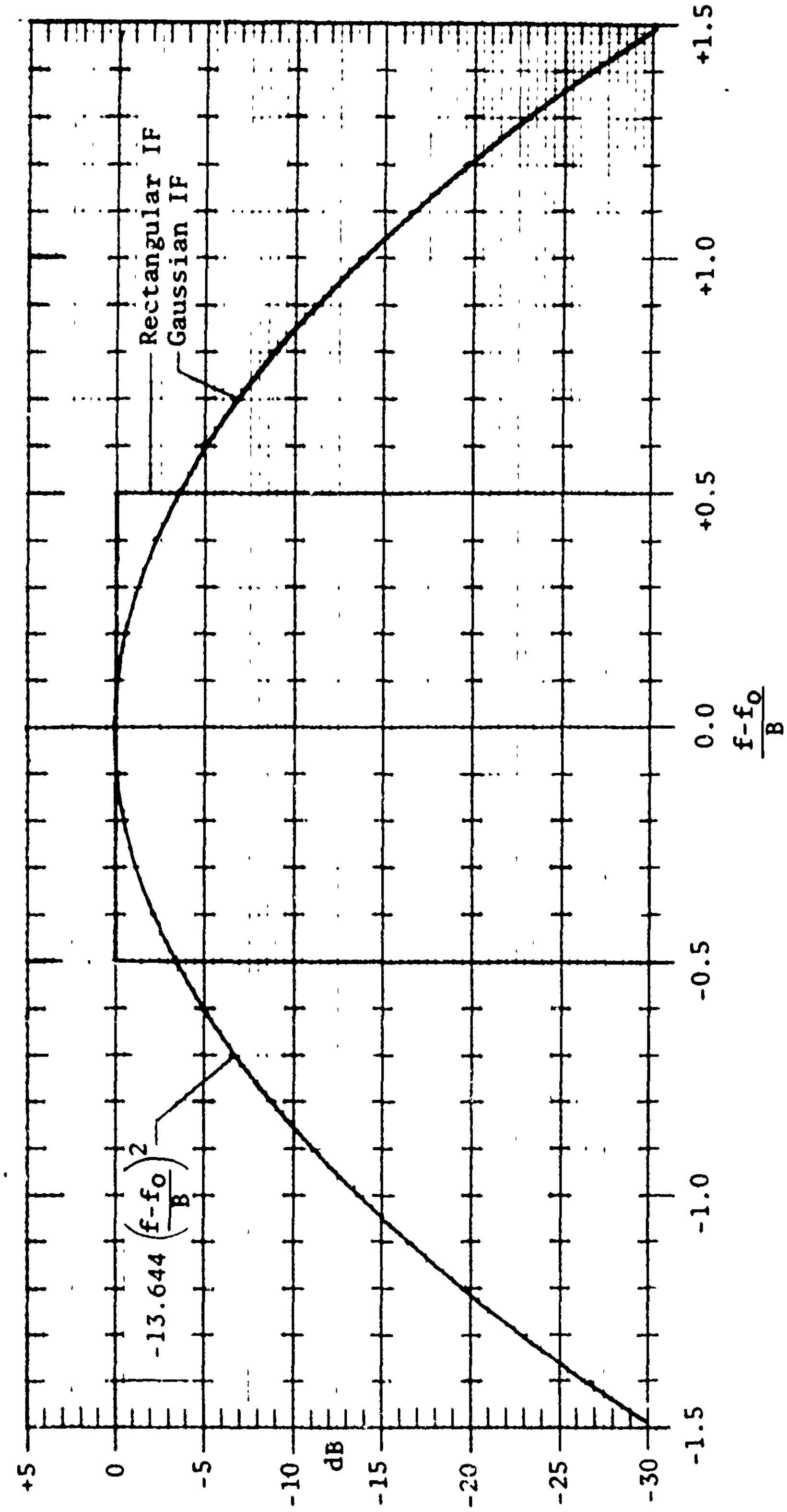
$$H(f) = e^{-(fo - f)^2 / 2\sigma^2}$$

σ = an arbitrary constant to allow the bandwidth to be varied

fo = center frequency of the IF

f = RF frequency

7.68 The power responses of the preceding IFs are tabulated on the next page. B is IF noise bandwidth, f is the frequency of measurement, and fo is the IF center frequency. Using the definition of noise band-



Gaussian/Rectangular IF
Power Response

Figure 78.1

width as defined in the noise figure section, using a change of variable, and taking advantage of the error function integral, it can be shown that the noise bandwidth of the Gaussian IF is $\sqrt{2\pi}\,\sigma$. Since the 3dB bandwidth is the range of frequencies between the half power points, equating the IF response to 1/2 and solving for the two frequencies of interest indicates a 3dB IF bandwidth of $2\sigma\sqrt{2 \ln 2}$.

7.69 In the noise quieting curve calculations, the IF bandwidth factor is 10 log B where B is the IF bandwidth. The dB error associated with assuming 3 dB bandwidth is therefore the following:

$$\text{error} = 10 \log\left(\frac{\text{3 dB bandwidth}}{\text{noise bandwidth}}\right)$$

$$= 10 \log\left(\frac{2\sigma\sqrt{2 \ln 2}}{\sigma\sqrt{2\pi}}\right) = -0.271 \text{ dB}$$

7.70 Therefore, the 3dB bandwidth is slightly more narrow than the noise bandwidth. The difference is negligible.

7.71 There are other considerations besides IF bandwidth. The shape of the IF power response effects the baseband slot morse performance of the M/W receiver. As can be noted by observing the tabulated slot noise values for Gaussian and rectangular IF responses, the IF response causes 0.2 to 0.3 dB difference in slot noise in the linear portion (region C) of the quieting curve. In the nonlinear transition region (region B) for low C/N ratios, IF response has a significant effect on the slot noise. There is as much as one dB difference in the slot noise due to the two different types of IF. From a practical point of view, however, this difference will be difficult to measure since other factors can significantly effect the slot noise in this region. The noise in this region is directly affected by the degree of limiting in the receiver for this region. Such information is seldom available. Errors in RF signal power measurement also cause the measured slot noise to change. In region B, an error of 1/4 dB in measurement of RF signal power can change the slot noise on the order of one dB. In the very low C/N ratio region (region A) where the input RF signal is insignificant (in terms of slot noise effect), IF shape again effects the slot noise by as much as one dB.

7.72 Unless the actual IF response is known to fit a Gaussian or rectangular IF response, it is suggested that noise measurements be compared to averaged slot noise values. This allows the IF shape to be disregarded but will introduce error on the order of 1/2 dB. This is an excellent compromise which, when viewed in terms of the practical constraints of noise measurement, will yield negligible error.

7.73 The other problem with regard to IF bandwidth is determining what the hypothetical performance of a radio would be with various parameters. The IF bandwidth of the actual radio might be unknown. The classical

approach to IF bandwidth estimation has been to use Carson's rule. Stated as a formula, the rule is:

$$B = 2\ f_{max} + 2\,\Delta f_{pk}$$

where

B = IF bandwidth (kHz)

f_{max} = maximum baseband frequency (kHz)

Δf_{pk} = peak deviation of the carrier away from rest (unmodulated) frequency

Smith has shown that Carson's rule can be converted to the following formula for wideband FM systems:

$$B = 2\ f_{max} + 2\ \Delta f_{rms}\ VF_{pk}$$

where

$\Delta f_{rms} = \Delta f_{/ch\ rms} \times VF_{rms}$

VF_{pk} = voltage peak factor (peak to rms voltage ratio) for white noise.

$\Delta f_{/ch\ rms}$ = rms per channel deviation (for a 0 dBmØ sine wave test tone)

= peak deviation for a peak voltage level equal to a DC level equivalent to 0 dBmØ rms

VF_{rms} = rms voltage level relative to a 0 dBmØ rms voltage level

= antilog (CCIR/20)

$= 10^{CCIR/20}$

$$\text{CCIR (noise loading ratio)} = \begin{cases} +2.6 + 2 \log n, & \text{12 to 59 channels} \\ -1 + 4 \log n, & \text{60 to 239 channels} \\ -15 + 10 \log n, & \text{240 or more channels} \end{cases}$$

n = number of channels

7.74 The main problems with this method is determining the voltage peak factor VFpk. There is no agreement among sources. As Goldman has pointed out, experimental values between 3.4 (+10.6 dB) and 4.5 (+13.1 dB) have been obtained for this factor by various sources. Lacking any general agreement, the Holbrook/Dixon values for peak factor were used. Comparing them with

the CCIR rms noise voltage values the chart of the VFpk factor (in dB) on the following page was derived.

7.75 Anuff and Lou at Bell Labs have developed a new formula for IF bandwidth. They have related required bandwidth of the IF to a required signal to intermodulation noise. Their formula is:

$$B = 2\,(1 + 0.065 \log (s/n)\,)\, f_{max} + \Delta f_{rms} \log (s/n)$$

where s/n = required signal to noise ratio (power ratio)

and other variables as previously defined.

or

$$B = 2\, f_{max}\,(1 + 0.0065\,(S/N)\,) + \frac{(S/N)}{10}\,\Delta f_{rms}$$

where

S/N = required signal to noise ratio in dB.

7.76 The above formula was developed with the assumption that the IF bandwidth is rectangular. Although the Anuff/Lou formula is useful, it trades the question of "What is the peak factor of white noise" for "What is the required signal to noise ratio of a rectangular IF which is equivalent to a practical IF." Lundquist, also of Bell Labs, suggests that if rectangular IF responses could be obtained, the S/N required would be 70 dB. Since in practice the filters are broader than the idealized rectangular ones, he suggests the use of 90 dB. Substituting 90 dB into the formula yields

$$B = 3.17\, f_{max} + 9.0\,\Delta f_{rms}$$

This formula bears considerable resemblance to Carson's rule. Both can be represented by the following general formula.

$$B = F1(\text{kHz}) + \Delta f_{/ch\ rms}(\text{kHz}) \times F2$$

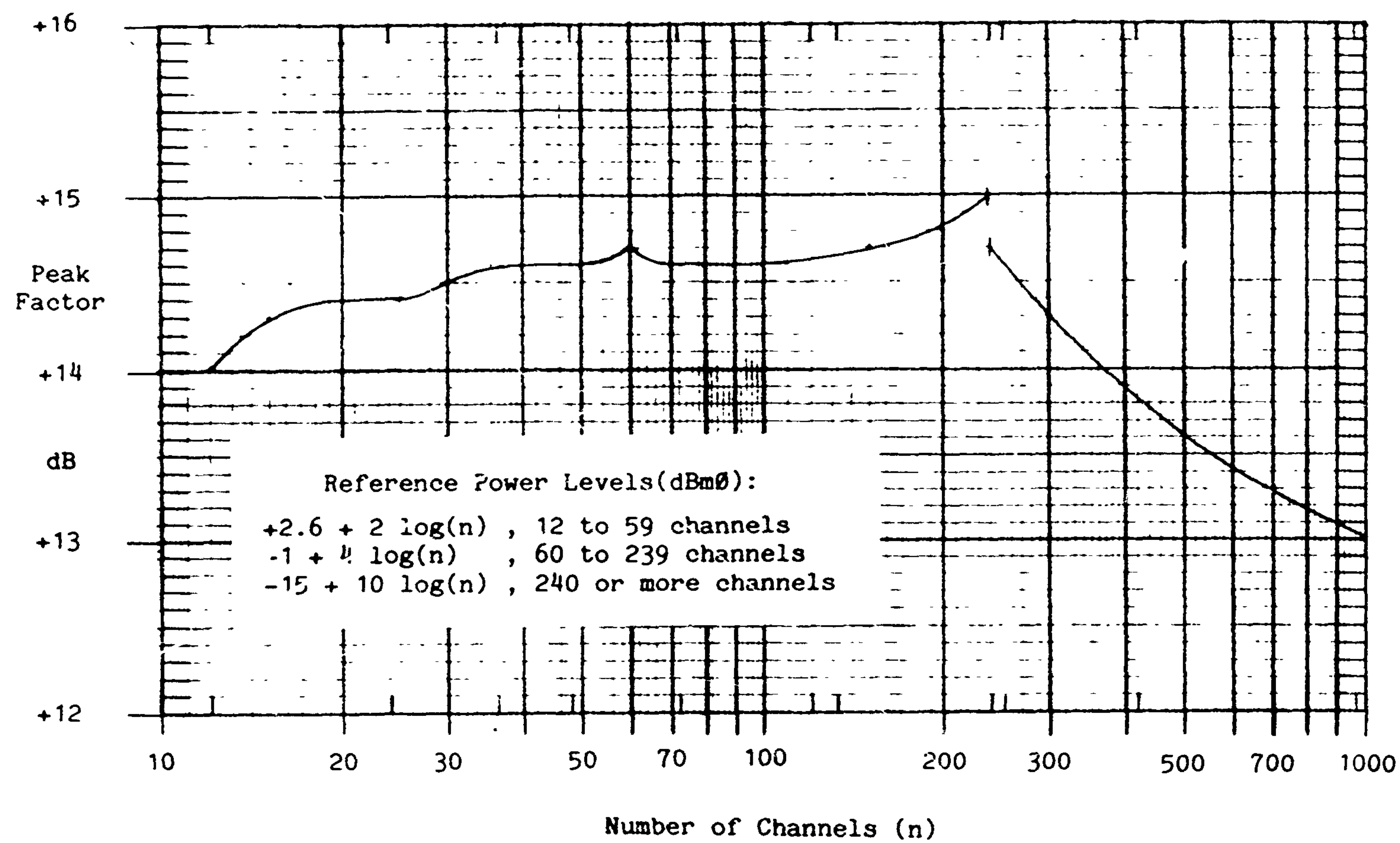
For Carson's rule

$$F1 = 2\, f_{max}$$

$$F2 = (2\,\Delta f_{/ch\ rms})\,(10^{PK/20})\,(10^{CCIR/20})$$

PK = peak factor(dB) from previous graph

$$= 10 \log (VFpk)$$



Holbrook/Dixon Composite Peak to RMS Voltage Ratios
(peak limiting assumed)

Figure 79

For the Anuff/Lou formula

$$F1 = (1 + 0.0065\ (S/N)\ )\ 2\ fmax$$

$$= 3.17\ fmax \text{ for } S/N = 90 \text{ dB}$$

$$F2 = 2\ \Delta f_{/ch\ rms}\ (10^{\ CCIR/20})\ ((S/N)/20)$$

$$= 9.0\ \Delta f_{/ch\ rms}\ (10^{CCIR/20}) \text{ for } S/N = 90 \text{ dB}$$

7.77 The preceding information has been summarized in the following chart. A/F means Anuff/Lou and C means Carson. CCIR noise loading (noise load ratio) is CCIR NLR. The following chart lists the required IF bandwidths used the two methods for the CCIR recommendations for per channel rms deviation.

7.78 It is not immediately obvious whether or not pre-emphasis on an FM transmitter will cause a change in required IF bandwidth. This question was answered by observing the IF bandwidth of a microwave transmitter with different baseband loading. The spectrums were compared with and without pre-emphasis. The results of this experiment are shown on the next two pages. The two baseband loading values present the normal range of spectrum shapes produced by microwave transmitters. The CCIR loading case represents a typical TROPO RF spectrum. The -2.89 dBm0 loading represents a typical LOS RF spectrum. Notice that although the RF spectrums are quite different in shape, the shapes do not change drastically with the use of pre-emphasis. Pre-emphasis can be disregarded as a factor in IF bandwidth estimation.

Noise Figure

7.79 Any meaningful evaluation of the noise performance of a microwave device must include an assessment of the noise that is added to a signal which is due just to the device itself. The noise performance of a device is dependent on the sources of internal noise, the bandwidth and gain of the device in all its responses, the nature of the signal, the linearity of the device, and the efficiency of the circuits. As Adler et al point out, "It is evident that no single number can describe completely how well a given receiver will perform in all kinds of systems." Never the less, it is necessary to have a number which characterizes the static (nonloaded baseband) noise performance of the device under nominal operating conditions. Noise figure is such a number. The basics of noise figures will be discussed in the next few pages. For those interested in more information, the references listed in the bibliography are suggested. Before discussing noise figures, however, it is necessary to do a little preliminary work.

7.80 Noise figure relates to the quasi-thermal noise introduced by the device into the signal path. Thermal noise is the type of noise generated

VF Channels (number)	f_{max} (kHz)	(dB) Peak Factor (VR)	(dBm0) CCIR NLR (VR)	F1 (kHz) F1	(C) (A/L)	F2 (none) F2
12	60.	14.0	4.758	120.000		17.336
		5.012	1.729	190.200		15.565
24	108.	14.4	5.360	216.000		19.456
		5.248	1.854	342.360		16.683
36	156.	14.6	5.713	312.000		20.733
		5.370	1.930	494.520		17.373
48	204.	14.6	5.962	408.000		21.338
		5.370	1.987	646.680		17.880
48	252.	14.6	5.962	504.000		21.338
		5.370	1.987	798.840		17.880
60	252.	14.7	6.113	504.000		21.961
		5.433	2.021	798.840		18.192
60	300.	14.7	6.113	600.000		21.961
		5.433	2.021	951.000		18.192
72	300.	14.6	6.429	600.000		22.516
		5.370	2.096	951.000		18.867
120	552.	14.6	7.317	1104.000		24.938
		5.370	2.322	1749.840		20.897
132	552.	14.7	7.482	1104.000		25.713
		5.433	2.367	1749.840		21.299
240	1052.	14.7	8.802	2104.000		29.932
		5.433	2.755	3334.840		24.794
252	1052.	14.6	9.014	2104.000		30.320
		5.370	2.823	3334.840		25.406
300	1300.	14.3	9.771	2600.000		31.959
		5.188	3.080	4121.000		27.721
420	1796.	13.8	11.232	3592.000		35.699
		4.898	3.644	5693.320		32.799
600	2660.	13.4	12.782	5320.000		40.748
		4.677	4.356	8432.200		39.203
900	4188.	13.1	14.542	8376.000		48.212
		4.519	5.335	13275.960		48.014
960	4028.	13.0	14.823	8056.000		49.223
		4.467	5.510	12768.760		49.588
1200	5564.	13.0	15.792	11128.000		55.033
		4.467	6.160	17637.880		55.441
1260	5564.	13.0	16.004	11128.000		56.392
		4.467	6.312	17637.880		56.810
1800	8204.	13.0	17.553	16408.000		67.401
		4.467	7.545	26006.680		67.901
2400	11404.	13.0	18.802	22808.000		77.828
		4.467	8.712	36150.680		78.406

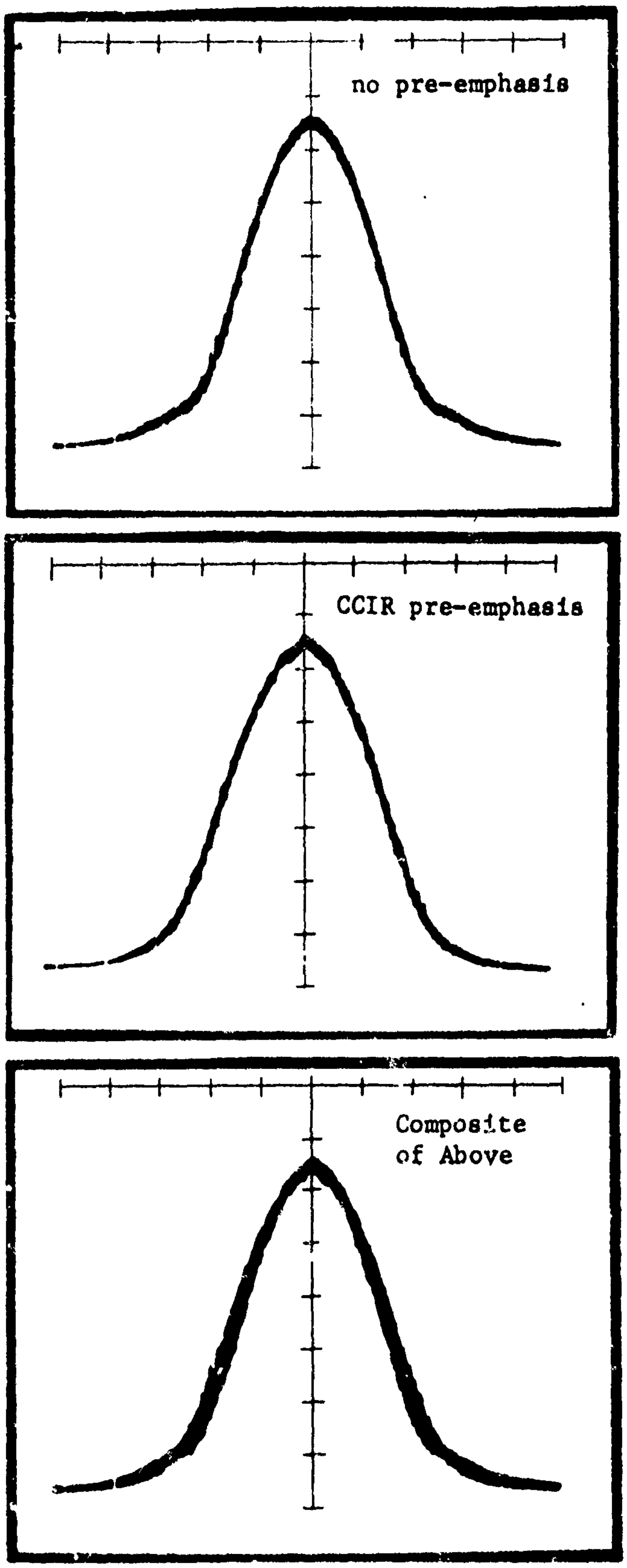
RF Bandwidth Factors

Table 33

VF Channels (number)	Baseband (kHz)	Deviation (kHz RMS/Channel)	Bandwidth Carson (kHz)	Anuff/Liou (kHz)
12	12-60	35	727	735
24	12-108	35	897	926
60	12-252	50	1602	1708
60	12-252	100	2700	2618
60	12-252	200	4896	4437
60	60-300	50	1698	1861
60	60-300	100	2796	2770
60	60-300	200	4992	4589
120	60-552	50	2351	2795
120	60-552	100	3598	3840
120	60-552	200	6092	5929
300	60-1300	200	8992	9665
600	60-2660	200	13470	16273
960	60-4028	200	17901	22686
1260	60-5564	140	19023	25591
1260	60-5564	200	22406	29000
1800	316-8204	140	25844	35513

RF Bandwidths for CCIR Deviation Recommendations

Table 34



Baseband Noise Spectrum:
12 to 252 kHz

CCIR Noise Loading
(+6.11 dBm0)

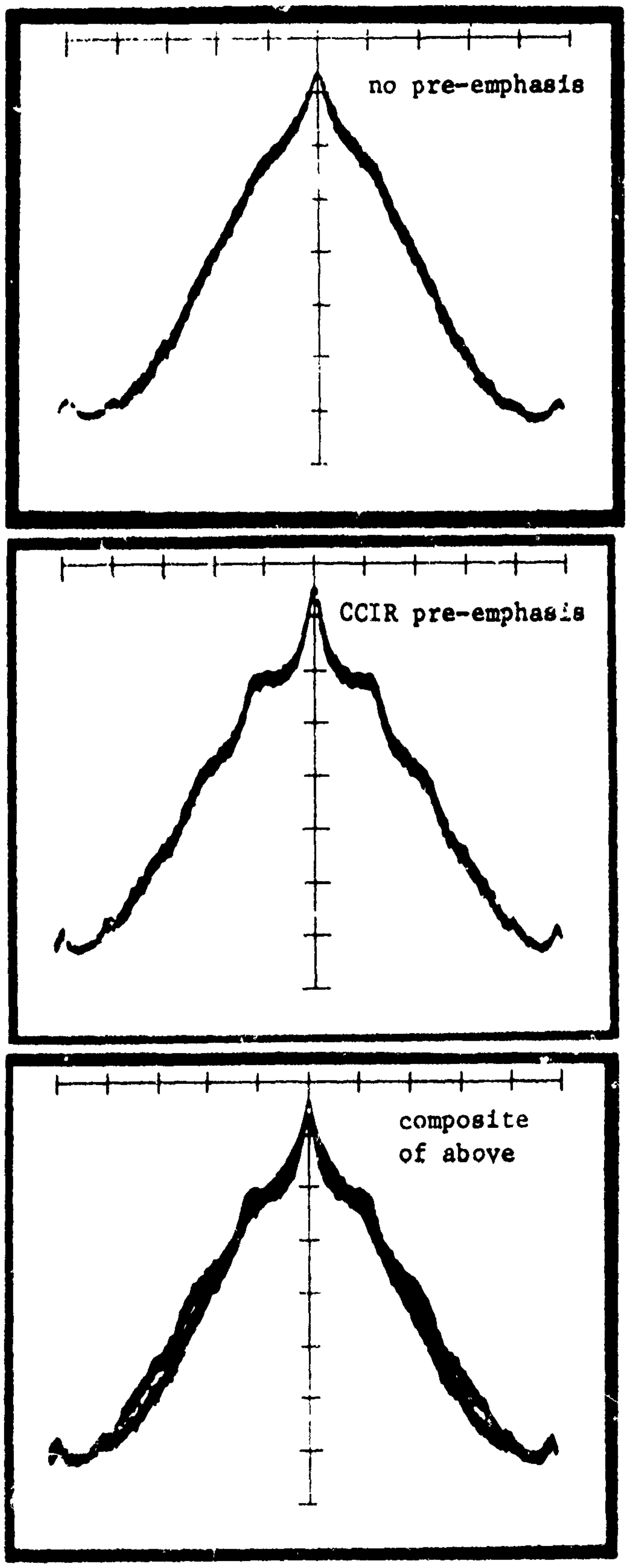
Vertical: 10 dB/Div.
Horizontal: 500 kHz/Div.

Reference RF power level is unmodulated carrier power level.

TROPO M/W Transmitter

FM RF Transmit Spectrum

Figure 80



Baseband Noise Spectrum:
12 to 252 kHz

-2.89 dBm0 Noise Loading

Vertical: 10 dB/Div.
Horizontal: 200 kHz/Div.

Reference RF power level is unmodulated carrier power level.

TROPO M/W Transmitter

FM RF Transmit Spectrum

Figure 81

by a resistor. The noise power generated by the resistor can be determined based on the value of the resistance and the absolute temperature of the resistor.

The mean squared voltage produced by the resistor is

$$\overline{en^2} = 4\ K\ T\ R \int_{f1}^{f2} \frac{hf/kt}{(hf/kt) - 1} \quad \frac{H(f)}{H(max)} \quad df$$

where $\sqrt{\overline{en^2}}$ = rms noise voltage

K = Boltzmann's constant

Teff = absolute (effective) temperature of resistor (degrees K)

f_1 to f_2 is the bandwidth of interest

R = value of the resistor (ohms)

h = Planck's constant

f = frequency (Hertz)

H(f) = power response (power out/power in) as a function of frequency of the device between the noise source (resistor) and the measurement device.

Hmax = maximum value of H(f) over the range f1 to f2.

7.81 As Vigg shows, for the frequencies and noise temperatures (noise figures) of interest to microwave engineers and technicians, the formula reduces to

$$\overline{en^2} = 4\ K\ Teff\ R \int_{f1}^{f2} \frac{H(f)}{Hmax}\ df$$

7.82 The integral portion of the above equation is the conventional definition (e.g., Davenport and Root) of noise bandwidth Bn of a network.

$$\overline{en^2} = 4\ K\ Teff\ Bn\ R$$

$$Bn = \int_{f1}^{f2} \frac{H(f)}{Hmax}\ df = \text{noise bandwidth}$$

7.83The above differs from most other definitions of noise bandwidth only in that the power response is indicated rather than the squared voltage response. Both methods are, of course, equivalent.

7.84 Consider the power delivered by a resistor R through a device with power frequency response H(f) to a load resistor RL. Considering a Thevinin's equivalent, the circuit can be idealized as shown on the next page.

The power delivered to RL is

$$P = \frac{(VRL)^2}{RL} = \frac{(\sqrt{\overline{en^2}}/2)^2}{RL} = \frac{\overline{en^2}}{4\ RL}$$

If we assume RL = R (matched source and load),

$$P = \frac{4\ K\ Teff\ Bn\ R}{4\ R} = K\ Teff\ Bn$$

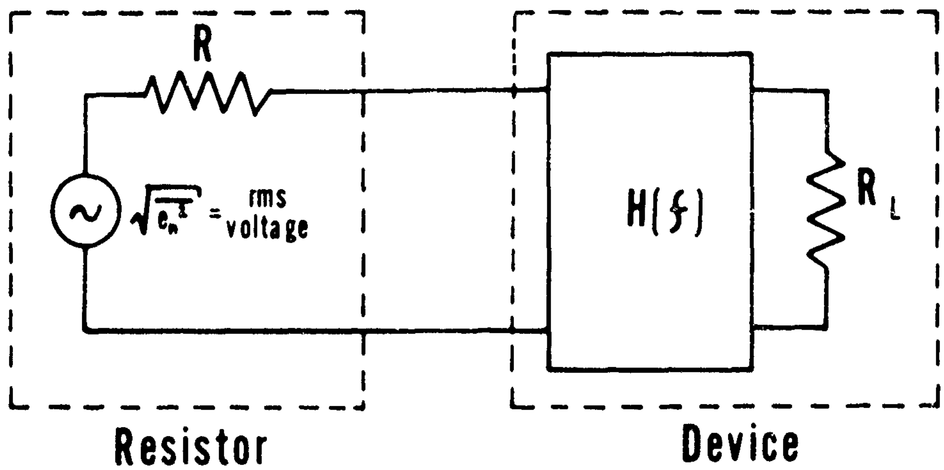
Therefore, the power delivered to load by a resistor under matched conditions is just K Teff Bn.

7.85 Now consider an idealized amplifier diagramed on the next page. The amplifier is assumed to be noiseless, but with power gain G and noise bandwidth Bn . A "ghost" resistor R is placed across the input terminals of the noiseless amplifier. The temperature of this resistor would normally be just room temperature (by convention assumed to be 290°K). This is called reference temperature Tref. This resistor at temperature Tref puts noise into the amplifier that would be introduced by the source resistor that drives the amplifier. It is assumed that the resistor R is the source impedance and that the source and the internal impedance of the amplifier are identical (matched impedances). The source and amplifier are assumed to have purely resistive impedances.

7.86 Any noise that an actual amplifier puts into the signal path is considered as an additional equivalent temperature Tequ (additional heating of the resistor) for the resistor. Tequ accounts for actual amplifier noise.

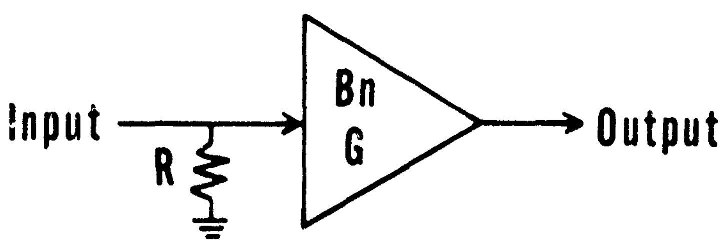
$$Teff = Tref + Tequ$$

7.87 If the actual amplifier were truly noiseless, Teq would be zero and the only noise out of the amplifier would be that of the driving source resistance R. Using a Tref equal to 290°K implies that the driving source in a microwave systems has no noise other than that produced physically by the signal sources. For microwave receivers, this source is a waveguide or coax transmission line preceeded by an antenna. Fortunately for analysis, the transmission line losses and terrestrial microwave receiver equivalent noise temperature (noise figures) are high enough that the excess noise introduced by the sky and earth can be disregarded.



Idealized Thermal Noise Power Transfer

Figure 82



Idealized Amplifier

Figure 83

7.88 Now let's consider noise figure. There is some disagreement as to just what is meant by the words noise figure and noise factor (e.g. see Mumford and Scheibe). For purposes of this report, a simplified view will be taken of noise figure. It will be assumed to be a spot (relatively narrow frequency) measurement and no special consideration will be made for multiple equipment responses. The measurement will be assumed to be made in the normal equipment configuration and environment.

7.89 The lower case nf will be used to represent noise figure as a power ratio and the upper case NF will be used to represent the decibel noise figure.

$$NF = 10 \log nf$$

7.90 Paraphasing Adler et al and Strum, the noise figure of a device is the ratio of noise power delivered by the output of a device to the noise power that the device would have delivered by the output if the device were noiseless and the input were terminated in a matched resistor at reference temperature Tref (290°K). A matched resistor is a resistor whose value is equal to the input resistance of the device.

Let

No = total noise power out of the device

NI = total noise power into a device due to input termination at Tref.

$$nf = \frac{No}{G\ NI} = \frac{SI\ No}{SI\ G\ NI}$$

Let

SI = input signal power

$$nf = \frac{SI/NI}{G\ SI/No} = \frac{SI/NI}{So/No}$$

where So = G SI = output signal power

Therefore, the noise figure is a direct measure of the degree to which a signal to noise ratio of an input signal is degraded.

$$nf = \frac{Teff}{Tref} = \frac{Tref+Tequ}{Tref} = 1 + \frac{Tequ}{Tref}$$

$$Tequ = (nf-1)\ Tref$$

Therefore, the equivalent noise temperature of a device is specified by a noise figure.

$$nf = \frac{No}{G\ NI} = \frac{G\ NI + Na}{G\ NI}$$

Where Na = noise added to the output by the device

$$Na = nf\ (G\ NI) - G\ NI = (nf-1)\ G\ NI$$

The noise added to the output of a device is specified by a noise figure.

Consider determining the noise figure of the cascaded devices diagramed on the next page.

G= G1 G2 = overall power gain (power ratio)
nf = overall noise figure (power ratio)
G1 = gain of first device (power ratio)
nf1 = noise figure of first device (power ratio)
G2 = gain of second device (power ratio)
nf2 = noise figure of second device (power ratio)
No1 = noise delivered by first device to second device

$$nf = \frac{No}{G\ NI}$$

$$No = G\ NI\ nf = No1\ G2 + Na2$$

where Na2 = noise added to output by the second device

$$No = No1\ G2 + (nf2 - 1)\ G2\ NI$$

$$= (G1\ NI\ nf1)\ G2 + (nf2 - 1)\ G2\ NI$$

$$= G1\ G2\ NI \left[nf1 + \frac{nf2-1}{G1} \right]$$

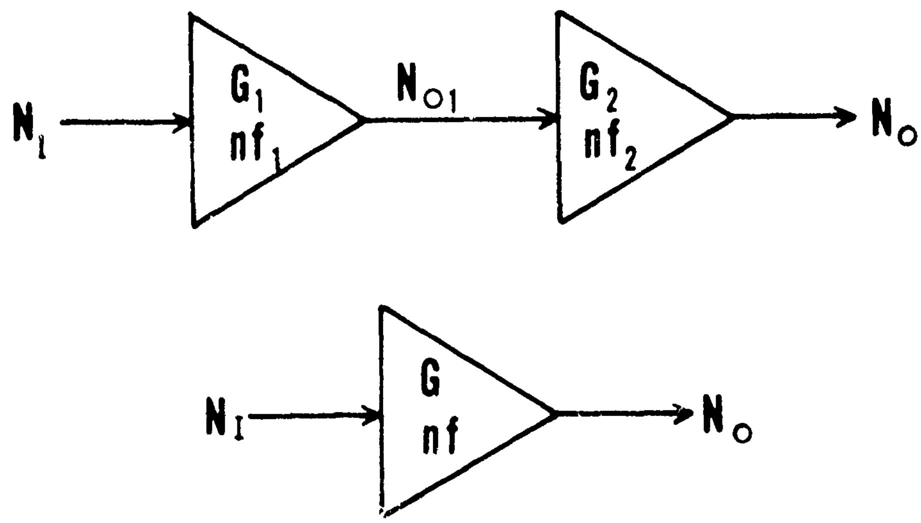
$$= G\ NI \left[nf1 + \frac{nf2-1}{G1} \right]$$

$$nf = \frac{No}{G\ NI} = nf1 + \frac{nf2-1}{G1}$$

In general,

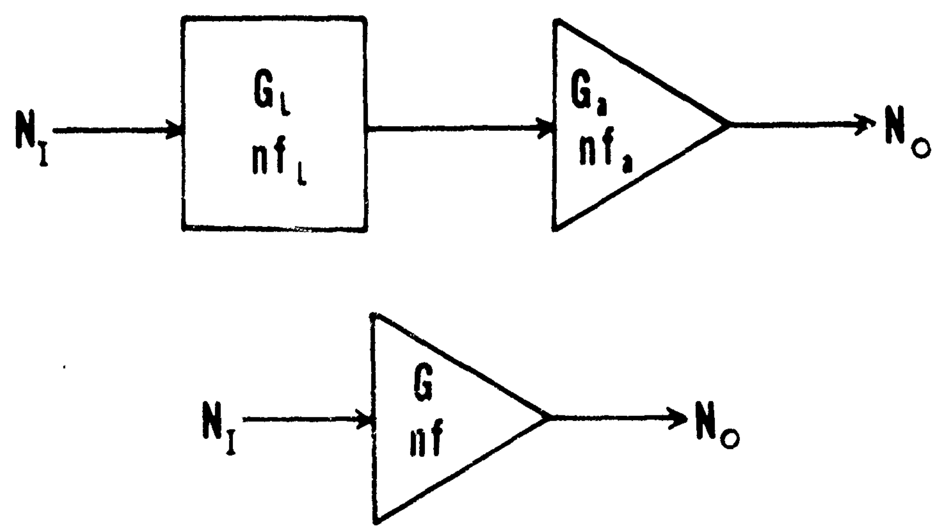
$$nf = nf1 + \frac{nf2-1}{G1} + \frac{nf3-1}{G1\ G2} + \ldots \frac{+nfn-1}{G1\ G2 \ldots Gn-1}$$

Therefore, given the noise figure of individual stages, it is possible to determine the overall noise figure of the device. It is assumed



Noise Figure of Cascaded Devices

Figure 84



Noise Figure of Lossy Network and Amplifier

Figure 85

that the stages are matched and that noise bandwidth shrinkage due to device cascading is insignificant (a good assumption for a few stages).

7.91 Consider determining the noise figure of a lossy device (e.g., cable, pads). If we assume matched devices and a purely resistive lossy device, the noise out of the device will be exactly the same as the noise into the device. The output of the device just looks like a resistor. Since the devices are matched, it looks exactly like the conceptual resistor at the device input. The signal out of the device is attenuated by the loss of the device.

$$nf = \frac{SI/NI}{So/No} = \frac{SI/NI}{SI\ G/NI} = \frac{1}{G}$$

For a lossy device, G is less than one.

$$NF = 10 \log (1/G) = \text{loss of device in dB}$$

7.92 The noise figure of a lossy device is exactly the loss associated with that device. It is worth noting that mismatches were not mentioned; matched devices was assumed. Any significant impedance mismatch will result in VSWR. This will result in increased signal loss which will cause increased noise figure. (See, for example, Livingston and Bechtold).

7.93 Consider an amplifier preceeded by a lossy network diagramed on the following page:

nf = overall noise figure
GL = gain of lossy network
(GL less than one,
10 log GL negative)
nfl = 1/GL
Ga = gain of amplifier
nfa = noise figure of amplifier

$$nf = nfl + \frac{nfa-1}{GL} = \frac{1}{GL} + \frac{nfa-1}{GL}$$

$$= \frac{1}{GL} (1 + nfa - 1) = (1/GL)\ nfa$$

$$NF = 10 \log ((1/GL)\ nfa)$$

$$= 10 \log (1/GL) + 10 \log (nfa)$$

$$= \text{loss of device (dB)} + NFa\ (dB)$$

This is an important practical result. It implies that the noise figure of a device under test is directly affected by the loss between it and

the input measurement port. This is the reason that, when predicting quieting curves, the RSL and NF must be specified for the same test point.

Specifically,

$$NFa = NF - L$$

where NF is a measured noise figure, L is the loss of the circuit (in dB) preceding the amplifier, and NFa is the actual noise figure of the amplifier.

7.94 In the preceding examples, we have assumed an amplifier (with a definite bandwidth Bn) so the situation would be easy to visualize. However, this amplifier can be replaced by any linear device as long as we don't violate any of the previous assumptions.

7.95 With the basics in mind, let's consider actual measurement. The most common method of noise figure measurement for microwave receivers is with a noise figure meter similar to the one described by Pastori. A typical test setup is diagramed on the next page.

7.96 The noise source is a device which can be used to deliver different amounts of noise into the microwave receiver. It terminates the input to the receiver and can be made to look like either a very hot or normal room temperature resistor.

7.97 Basically, the noise figure meter measures the noise out of the receiver with the noise source on (resistor "hot") and then measures the noise out of the receiver with the noise source off (resistor "cold"). The meter actually measures a Y factor.

$$Y = \frac{Nosh}{Nosc}$$

where Nosh = noise power out, source hot
and Nosc = noise power out, source cold

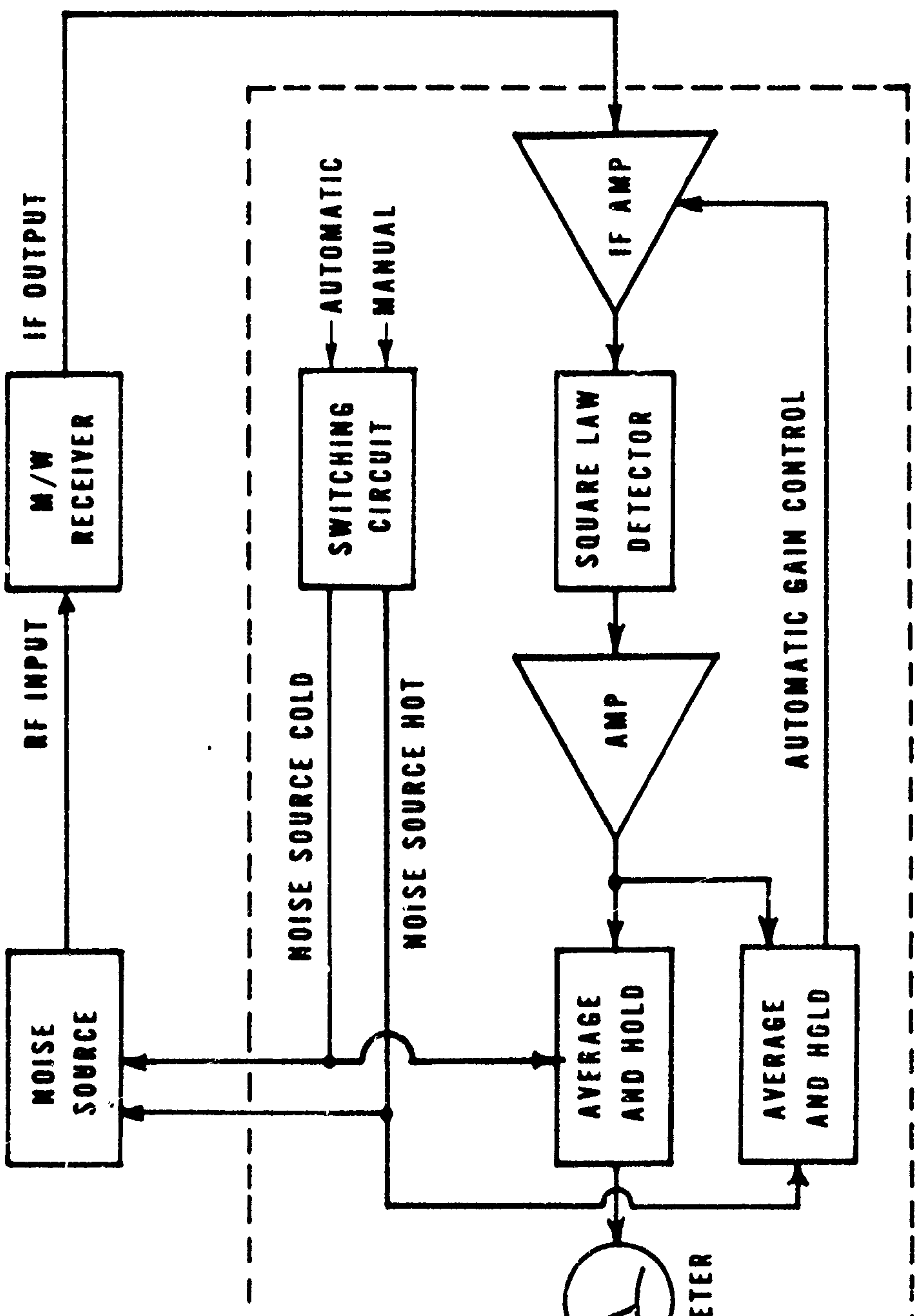
$$Y = \frac{Tsh + Ta}{Tsc + Ta}$$

where Tsh = source temperature with source "hot"
Tsc = source temperature with source "cold"
Ta = noise temperature of the receiver

The cold temperature is designed to approximate 290°K.

From the previous work on noise figures, it follows that

$$Y = \frac{Tsh + Ta}{Tsc + Ta} = \frac{Tsh + (nf-1)\ Tref}{Tsc + (nf-1)\ Tref}$$



NOISE FIGURE METER (SIMPLIFIED DIAGRAM)

Figure 86

Therefore

$$nf = \frac{(Tsh - Tref) - Y\,(Tsc - Tref)}{(Y-1)\ Tref}$$

$$= \frac{Tsh - Tref}{Tref} \quad \frac{1}{Y-1}$$

if $Tsc = Tref = 290^{o}K$

$$NF(dB) = 10 \log\left(\frac{Tsh - 290^{o}}{290^{o}}\right) - 10 \log (Y-1)$$

since $10 \log\left(\frac{Tsh - 290^{o}}{290^{o}}\right)$ is defined

as the "excess noise" of the source

$$NF(dB) = \text{excess noise(dB)} - 10 \log (Y-1)$$

7.98 The noise figure meter actually measures the Y factor, but the meter is calibrated to read noise figure in dB.

7.99 There are several possible sources of error in the measurement of noise figure. The most obvious error is the error associated with the "excess noise" of the noise source. When a noise figure meter is designed, it is designed for a nominal noise source excess noise. The noise figure meter is used to measure the noise figure of devices with inputs in the frequency range of a few hundred megahertz to several gigahertz. This requires different noise sources, each with different actual excess noises. Also, the excess noise of a given noise source is not constant over its normal frequency range. Therefore, it is necessary to add a correction factor to the noise figure reading to account for the difference between the actual source excess noise and the nominal excess noise. The correction is to algebraicly add the difference between the nominal excess noise and the actual excess noise to the measured noise figure. Although this source of error is often overlooked, it can cause over one dB error in the measurement. Another cause of error is failure to correct for loss between the noise source and the input to the device being tested (e.g., cable loss). Any errors in this loss cause a dB for dB error in the measurement.

7.100 Greene mentions several other ways to alter the measured noise figure of a device. One way to introduce excess gain between the IF output of the receiver and the input to the noise figure. This is easy to do because, in the field, wide band amplifiers with potentially very high gain are used to make up for losses in the system so that the meter can be calibrated during initial alignment. Excess gain can cause two undesirable effects. First, it may saturate the IF amplifier in the meter. This will drive the AGC loop to saturation and cause errors

in the measurement of Nosc. The saturated IF (and perhaps video) amp can cause clipping of the noise, changing its peak factor, and therefore, change the calibration factor for the square law detector. This causes the meter reading to be in error.

7.101 A similar problem is RFI. If any approximately 70 MHz signal gets into the meter, the measurement will be affected. In extreme cases, it can saturate the IF amplifier.

7.102 Another way to cause errors is to put a precision pad ahead of a low noise figure device and then correct for the pad loss. If the pad has high enough attenuation, it is possible to completely swamp the device noise figure due to uncertainty in the actual absolute loss of the pad and its source and device impedance match (VSWR). This method is the best way to alter noise figure measures virtually at will. Negative noise figures can be obtained, if desired.

7.103 As Okean and Lombardo observe, it is possible to make some microwave receivers appear to have noise figures lower than the "operational" noise figure. What will occur in practice depends on several factors such as flatness of noise source, source/receiver broadband impedance match, and down converter match at image frequencies. Basically, however, the procedure is to measure the noise figure of a receiver without its normal preselector or to design the receiver with an excessively wide preselector. This will allow noise to be introduced into the receiver at the image frequency when the source is hot. Under worst case conditions, the measured noise figure could be improved 3 dB. Obviously the normal operation of the device has not been improved by doing this. The noise figure of a mixer measured through a preselector which excludes noise at the image frequency is called a single sideband noise figure. The noise figure of a mixer measured by a method which allows both the noise at the desired frequency and the image frequency to be down converted to IF is called a double sideband noise figure.

7.104 Wait mentions additional factors of error such as system gain changes (due to gain instability or nonlinear operation) between noise source hot and cold conditions, impedance mismatch uncertainty, Tsh and Tsc uncertainty, and uncertainty in measurement system noise. Wait's NBS Technical Note 640 provides an excellent set of charts for estimating noise figure error.

7.105 For most line of sight (LOS) terrestrial microwave (M/W) receivers, the RF signal signal passes through an RF filter, transmission line, and is then applied to a crystal mixer (down converter). A preamplifier (preIF amp) immediately follows the mixer to avoid degrading the noise figure of the receiver by post mixer cable losses. For the typical LOS receiver configuration just mentioned, the overall noise figure of the receiver is determined by the RF losses prior to the mixer and by the noise figure of the mixer itself. The gain of the pre IF amplifier makes the noise figure of the receiver virtually independent of following

stages (unless abnormally high losses develop in following stages). The RF losses prior to the mixer are simple to understand and measure. The causes of noise figure of a mixer are not so obvious. As Neidert, Pritchard, and Roberts mention, the noise temperature (noise figure) of a crystal mixer depends c the inherent crystal noise temperature, the method of mounting the crystals and crystal bias, source impedance, and conversion efficiency (power out/power in). Conversion efficiency is a function of drive level and the mixer's load (preIF amp) impedance at the input/drive sum and difference frequencies. Edwards gives the formula:

$$nf = nfa + L\,(nfb - 1)$$

where

nf = overall noise figure of mixer
nfa = noise figure of diodes
nfb = noise figure of preIF amplifier
L = conversion loss (power ratio) of mixer

7.106 From a practical point of view, the most important parameter of a properly designed M/W receiver is the local oscillator drive level into the mixer. If it is too great, intermodulation performance of the receiver is degraded. If the drive is too low, the noise figure of the receiver is degraded due to excessive conversion loss.

7.107 For the typical tropospheric scatter (TROPO) M/W receiver, a low noise figure is placed in front of the receiver mixer. If the gain of this amplifier is high, the overall noise figure of the receiver will be limited primarily by the RF losses in front of the RF amplifier and the noise figure of the amplifier itself.

7.108 It should be noted that, although the overall noise figure of a M/W receiver is normally limited by the RF losses and first active devices in the receive, excessive losses in other portions of the receiver can affect overall noise figure. The excessive loss will degrade the noise figure directly and will also greatly magnify the effect on overall noise figure of the noise figure of the next stage following the loss.

7.109 Occasionally it is desirable to determine the noise figure of a M/W receiver but a noise figure meter is not available. If the equipment parameters are known (and have been verified by actual measurement), it is possible to determine the noise figure using the quieting curve formula. Rewriting the quieting curve prediction formula yields the following:

$$NF(dB) = +139.1 + N(dBm\emptyset) + P(dB) - 20\log f(kHz) + RSL(dBm) + 20\log \Delta f_{ch\,rms(kHz)}$$

7.110 The above formula is for an averaged response and assumes operation in the linear portion of the quieting curve (region C) and, of course, a reasonably flat symmetric IF frequency response with the receive carrier centered in the bandpass of the receiver IF amplifier. The symbols are the same as those used in the other sections of this report.

7.111 A known power level RF signal is applied to the input of the receiver and the baseband slot noise for that signal level (simulated RSL) is measured with a 3.1 kHz wide (3 dB bandwidth) frequency selective voltmeter. Using that information and the receiver characteristics, the above formula yields overall operational noise figure.

7.112 To yield accurate results will the above formula, it is imperative that the required parameters be known accurately. The formula method is relatively insensative to slight errors in 3 dB IF bandwidth measurements. However, it is quite sensitive to errors in noise measurement and RSL. These errors effect the result on a dB for dB basis. If de-emphasis is used in the receiver, the frequency response of the de-emphasis network must be known. In practice, the de-emphasis can be significantly different than the idealized frequency response.

7.113 There is one other source of error which can be quite significant. That is the actual TLP of the receiver baseband. Since the slot noise will actually be measured in dBm, conversion of the slot noise to a dBmØ value requires an accurate determination of TLP. To double check the alignment of the receiver, it is recommended that a signal generator be FM modulated with a sine wave. Taking advantage of the carrier drop out properties of the modulated RF signal, it is possible (knowing the deviation sensativity of the receiver) to produce a frequency deviation which will produce a known level in the baseband. For example, if a M/W FM receiver has been tuned to a per channel rms deviation of 140 kHz, a M/W FM transmitter with first carrier dropout produced by a 82.3 kHz test tone will produce, by definition, a 0 dBmØ 82.3 kHz test tone in the baseband of the receiver (disregarding de-emphasis). If LOS link measurements are being made, the far end transmitter can be used as the signal source. Carrier drop out (with the far end transmit pilot disabled) can be observed at the receive end using a spectrum analyzer at an RF test point or at the IF output of the pre IF amplifier. Use of this double check on the receiver alignment is highly recommended if the above formula is to be used to determine receiver noise figure.

8. A Digital FM System.

8.1 The preceding analysis has been limited to analog FDM M/W wideband FM transmission. The analysis can be applied to many other types of FM transmission systems. As an example, the analysis of a current wideband digital FM transmission system will be considered. It is not the object of this report to investigate digital techniques. A method of rough analysis will be briefly outlined.

8.2 The Air Force, in its Digital European Backbone (DEB) wideband communication upgrade, is converting from analog to digital FM M/W transmission. In this transmission system a high speed digital signal (three level partial response) characterized by three distinct voltage levels is transmitted over a conventional analog FM M/W transmission system. Using a simple algorithm, the receive digital multiplexer converts one of three received voltage levels into a high speed binary digital signal. Noise degradation of the received three level signal by the M/W terminal can cause errors in the reconstructed digital signal.

8.3 For a well randomized three level digital signal, it can be shown that for an optimally adjusted (for Gaussian noise rejection) digital multiplex receiver, the probability of an erroneous voltage level decision by the demultiplexer is

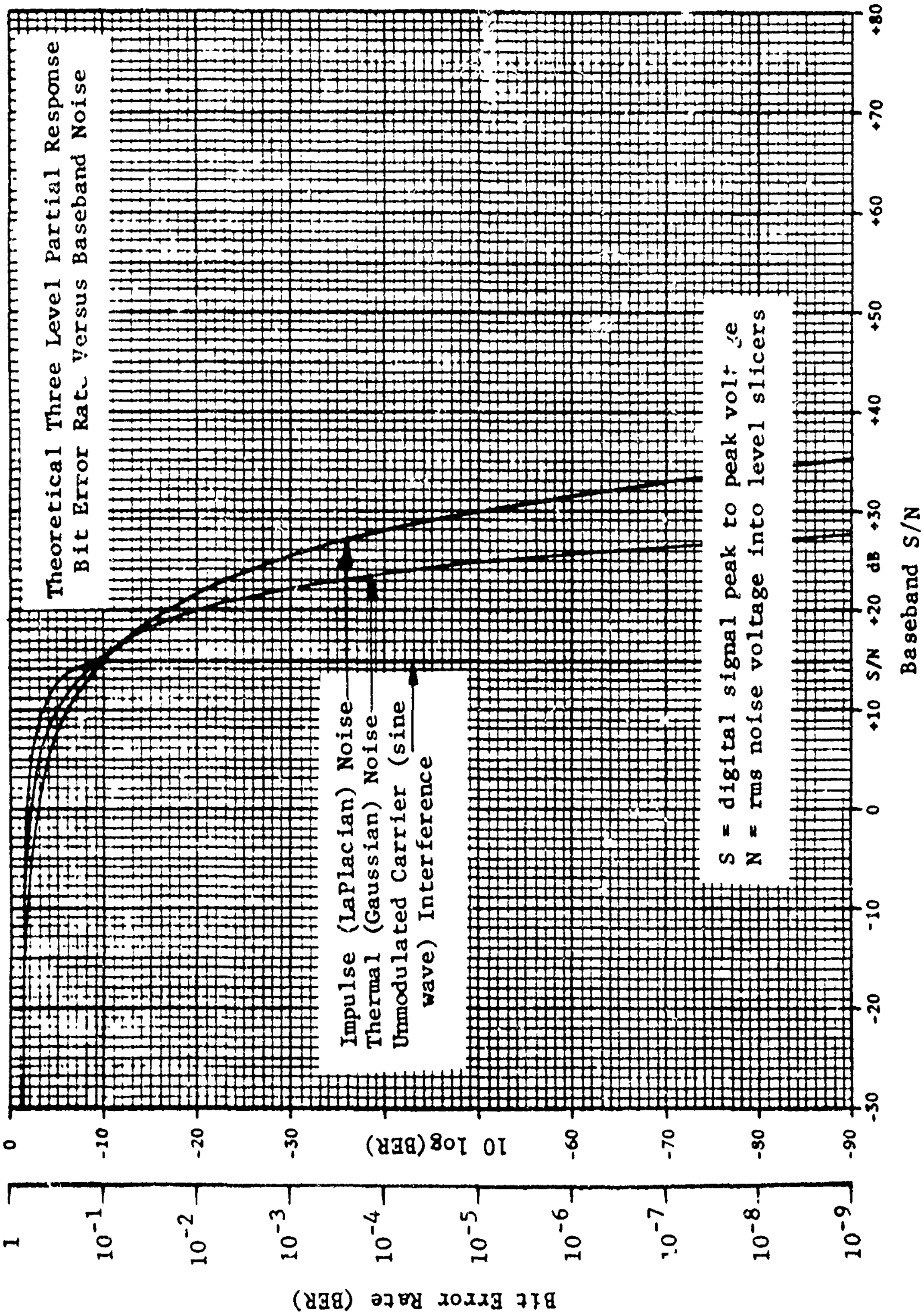
$$(3/2)\ F((S/N)/4)$$

where F(x) is the cummulative distribution function of the receive baseband noise, S is the baseband signal peak to peak voltage level, and N is the baseband noise rms voltage level. If the error extension of the digital signal reconstruction algorithm is disregarded, the bit error rate (BER) of the reconstructed digital signal becomes

$$BER = (3/2)\ F((S/N)/4)$$

Using the above relationship, theoretical BER curves were plotted on the next page for three different types of possible baseband noise. Following that page is a chart showing the comparison of the above formula and data collected by Hanlon and Reuter using white Gaussian noise. Since their experiment included a two feedband descrambler circuit, their BER results were divided by three prior to plotting to delete the effect of the descrambling circuit error extension.

8.4 For an FM receiver, the baseband noise will normally be Gaussian thermal noise. The noise spectrum will not be flat (white) but whiteness (or any other spectral distribution, for that matter) is not a necessary assumption for this analysis. If the noise is Gaussian, the cummulative distribution function is one half the complementary error function. For arguments greater than three, the following approximation yields less than ten percent error.



Theoretical Three Level Partial Response
Bit Error Rate Versus Baseband Noise
Impulse (LaPlacian) Noise
Thermal (Gaussian) Noise
Unmodulated Carrier (sine wave) Interference
S = digital signal peak to peak voltage
N = rms noise voltage into level slicers
-30
-20
-10
0
+10
S/N
+20
dB
+30
+40
+50
+60
+70
+80
Baseband S/N
0
-10
-20
-30
10 log(BER)
-60
-70
-80
-90
1
10^{-1}
10^{-2}
10^{-3}
10^{-4}
10^{-5}
10^{-6}
10^{-7}
10^{-8}
10^{-9}
Bit Error Rate (BER)

Figure 87

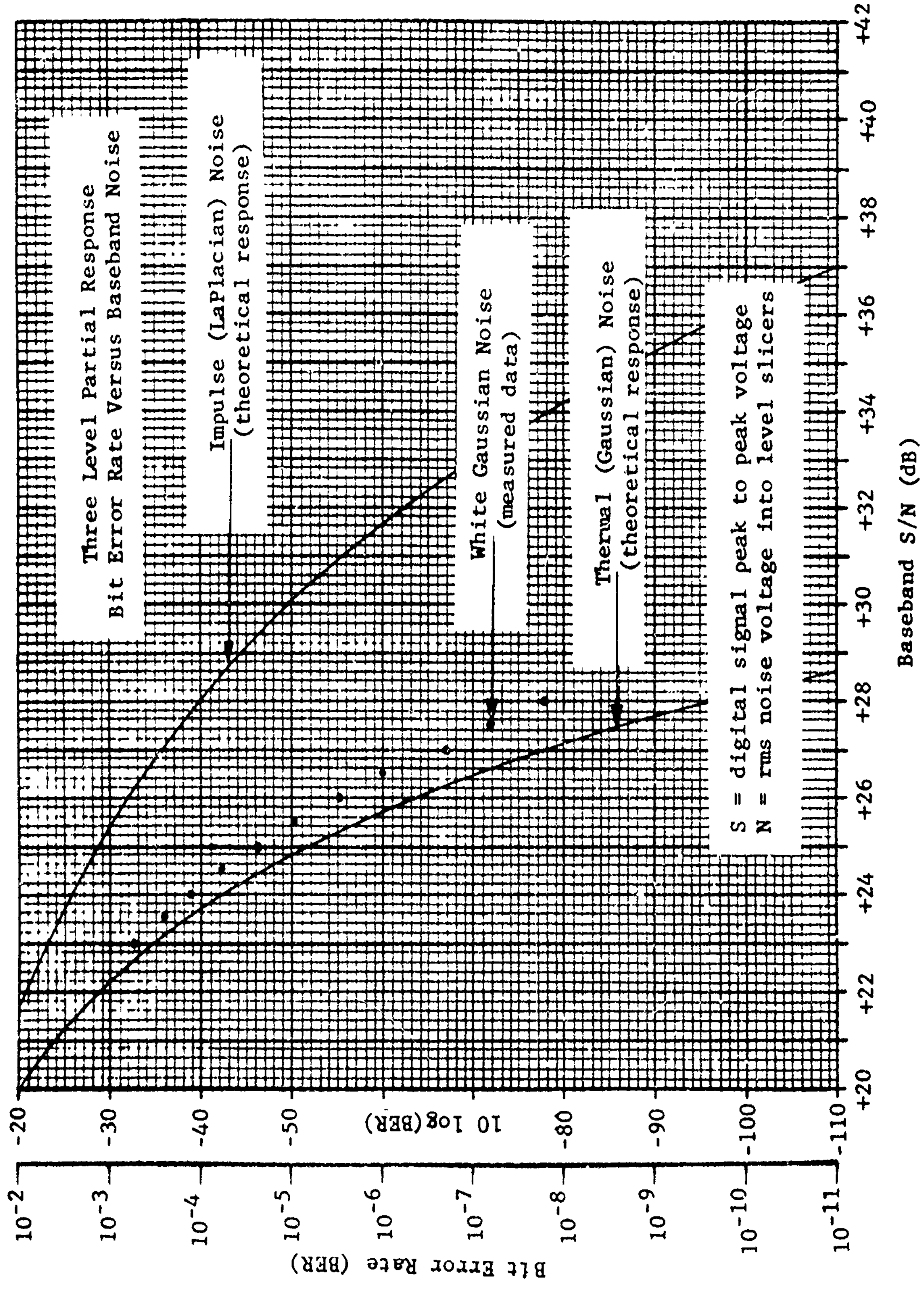

Three Level Partial Response
Bit Error Rate Versus Baseband Noise
Impulse (LaPlacian) Noise
(theoretical response)
White Gaussian Noise
(measured data)
Thermal (Gaussian) Noise
(theoretical response)
S = digital signal peak to peak voltage
N = rms noise voltage into level slicers
Baseband S/N (dB)
+20 +22 +24 +26 +28 +30 +32 +34 +36 +38 +40 +42
10 log(BER)
-20 -30 -40 -50 -80 -90 -100 -110
Bit Error Rate (BER)
10^{-2} 10^{-3} 10^{-4} 10^{-5} 10^{-6} 10^{-7} 10^{-8} 10^{-9} 10^{-10} 10^{-11}

Figure 88

$$F(x) = 1/2 \operatorname{cerf}(x) \cong \frac{e^{-x^2/2}}{\sqrt{2\pi}\,(|x|)}$$

8.5 For the DEB digital multiplex equipment, the receive circuitry has a six cell descrambler which, for error rates less than about 10^{-3}, has a constant error extension factor of three. Therefore, the following formula estimates the BER for the DEB equipment.

$$BER = (7.181/VR)\ e^{-(0.03125)VR^2}$$

where VR = signal peak to peak voltage to noise rms voltage (passed by the receive multiplexer filter) ratio

8.6 To determine the voltage ratio VR to apply to the above formula, the baseband signal and noise values are determined in dBm quantities, the values are subtracted and the result converted to a voltage ratio.

8.7 The baseband signal has a peak to peak value (at the combined baseband 75 ohm output) of 1 volt peak to peak. Converting this to a pseudo-dBm value results in a signal peak to peak "power level" of +11.3 dBm. The use of this dBm value is unconventional but the result obtained is the same as would be obtained by working directly in voltage ratios.

8.8 The 8 port digital multiplexer receive filter frequency response is three dB down at 4.71 mHz. Using the previously mentioned special case for total baseband noise yields the following power level for the previously mentioned LOS M/W FM receiver (with no de-emphasis and sharp baseband low pass filter).

$$\underline{N}(dBm0) = -148.7 + 30 \log f_{max} + NF - 20 \log \Delta f_{/ch\ rms} - RSL$$

$$= -148.7 + 30 \log (4710) + 8.1 - 20 \log (140) - RSL$$

$$= -73.7 - RSL$$

8.9 Since the FM terminal uses switching combining, no combiner improvement is used. However, multiplying the digital multiplex receive filter power response by f^2 and performing numerical integration indicates that 1.7 dB more noise than is predicted by the preceding low pass filter assumption will be produced by the FM receiver. The TLP of the FM terminal output is -21.8 dB. Therefore, the total noise power from the receiver is

$$\underline{N}(dBm) = -73.3 + 1.7 - RSL - 21.8$$

$$= -93.4 - RSL\ (dBm)$$

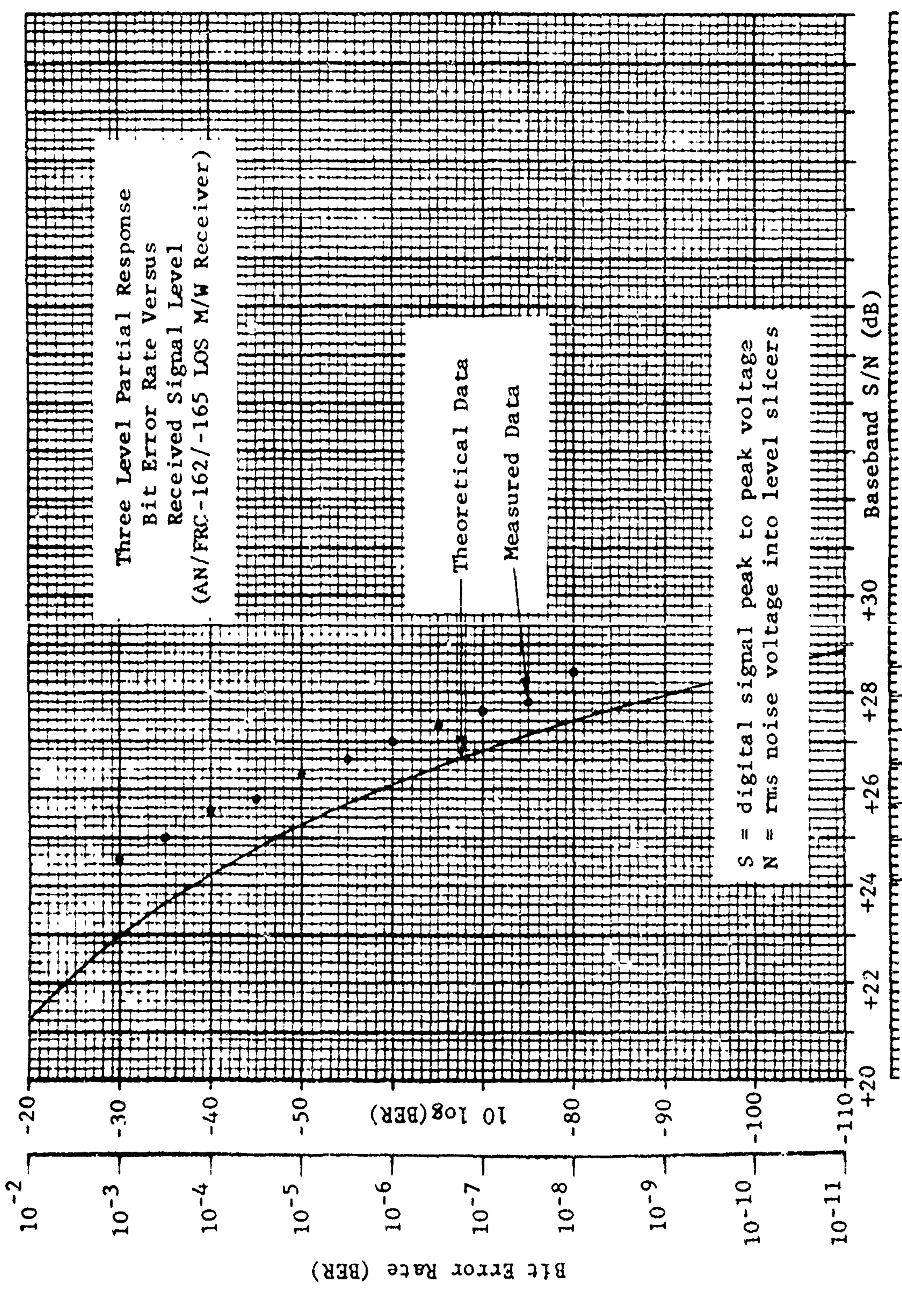
8.10 In addition to the radio thermal noise, the order wire and carrier and radio pilot signals increase the total noise power. The total pilot and orderwire power into the digital multiplexer (including receive multiplexer filter response) is about -21.1 dBm. This power can be added on an rms basis to the thermal noise produced by the receiver. This procedure is simplistic. The introduction of non-Gaussian noise changes the probability density function and cummulative distribution function for the noise process. Use of rms power addition will yield pessimistic signal to noise ratios as the non-Gaussian noise becomes dominate. However, this method will be used for ease of computation.

S/N (dB) = +11.3 dBm - N(dBm)

N(dBm) = rms sum of radio thermal noise ($\underline{N}$) and radio nonthermal noise

$$VR = \text{antilog } ((S/N)/20) = 10^{(S/N)/20}$$

8.11 The following page shows the theoretical BER curve produced using the above procedure. Also plotted is actual data obtained by J.R. Hammer using an eight port digital three level partial responses multiplexer and an LOS M/W receiver similar to the one used in the rest of this report. In addition to the previous assumptions, it should be kept in mind that the bit error rate estimates are only appropriate over the RSL range were baseband signal suppression and FM noise threshold are not significant effects and baseband noise is predominately radio thermal (f^2) noise.



Three Level Partial Respose
Bit Error Rate Versus
Received Signal Level
(AN/FRC-162/-165 LOS M/W Receiver)
Theoretical Data
Measured Data
S = digital signal peak to peak voltage
N = rms noise voltage into level slicers
Baseband S/N (dB)
RSL
dBm
+20
+22
+24
+26
+28
+30
-80
-75
10 log(BER)
-20
-30
-40
-50
-60
-70
-80
-90
-100
-110
Bit Error Rate (BER)
10^{-2}
10^{-3}
10^{-4}
10^{-5}
10^{-6}
10^{-7}
10^{-8}
10^{-9}
10^{-10}
10^{-11}

Figure 89

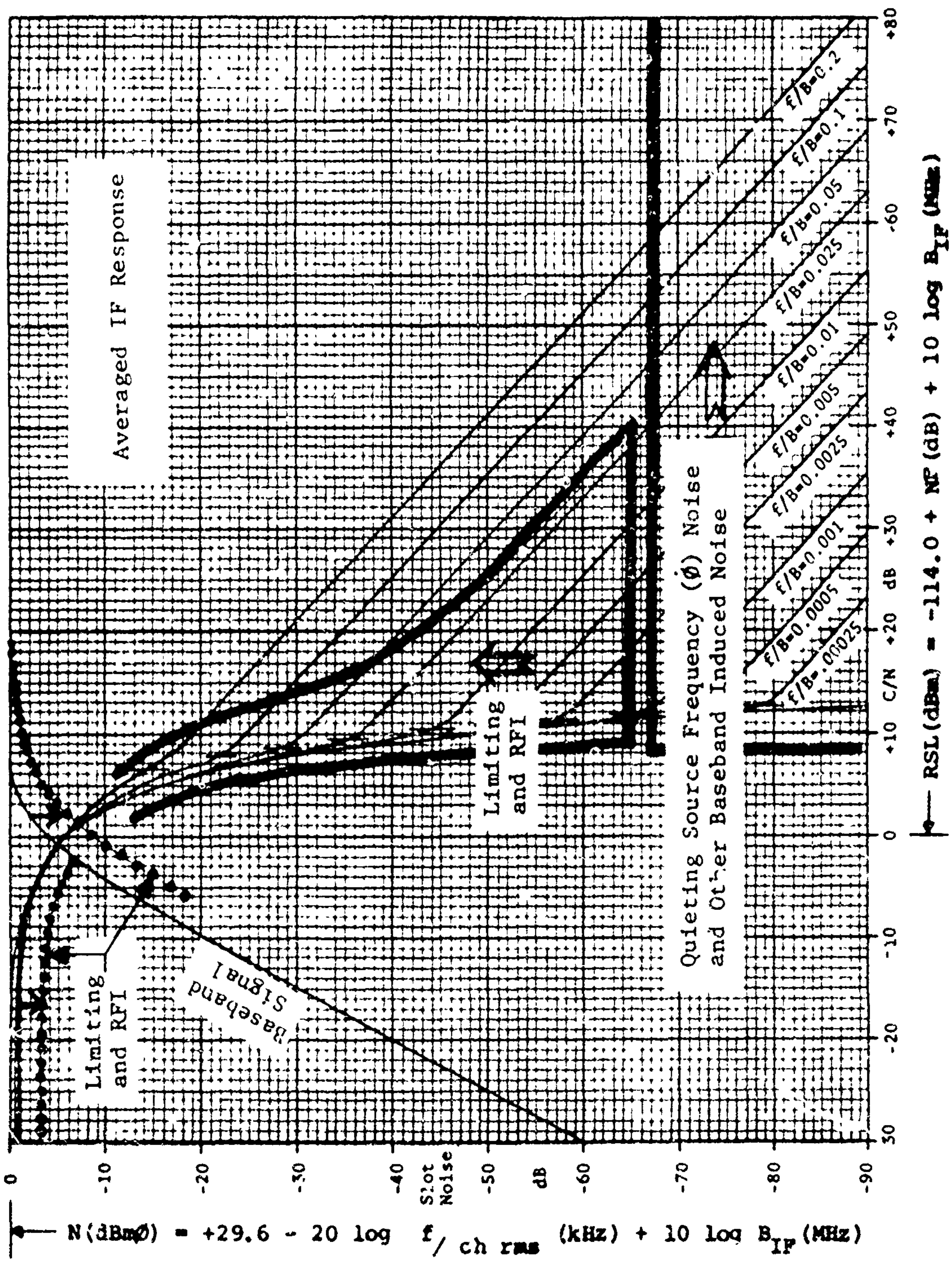
9. Conclusions

9.1 The reliability and performance of a M/W terminal is a significant factor in the reliability and performance of long distance telephone and low speed data communication. Baseband signal stability and telephone channel rms and impulse noise levels are primary performance factors directly influenced by FM M/W terminal performance.

9.2 The effect of nonideal limiting in an FM M/W receiver is significant. Soft limiting is most likely to occur for weak RSLs. Fortunately, a method is available for identifying lack of limiting in this case. Using a spectrum analyzer with storage CRT, baseband slot noise voltage or power versus frequency can be observed for various RSLs near 0 dB C/N. The receiver baseband de-emphasis network or low pass filter should be defeated to accomplish this test. The results can be compared with the two appropriate baseband slot noise charts for hard and no limiting. For FM receivers with broad IF frequency responses (approximately Gaussian), the baseband noise will be similar to one of the graphs. If the receiver's IF frequency response is sharp (approximately rectangular), the baseband slot noise will be similar to the graphs except the noise will peak at lower baseband frequencies and the noise will fall off faster with increasing baseband frequency.

9.3 The baseband signal of a M/W FM receiver falls off (is suppressed) rapidly for weak RSLs (low C/Ns). This suppression is readily predicted. The suppression occurs for stronger RSLs and occurs more rapidly with decreasing RSL as M/W receiver goes soft or if the M/W receiver experiences RFI. The degrading effect is greater for RFI with a relatively large frequency off set relative to the desired receive RF signal frequency.

9.4 The theoretical M/W FM receiver multiplexed telephone channel (slot) noise versus RSL is readily predicted. The baseband slot noise is increased for a given RSL as receiver limiting goes soft, if receiver noise figure is degraded, or if RFI is present. As limiting becomes soft, slot noise increases and the transition from region C to region B becomes less sharp. However, in region A, slot noise is reduced. Baseband noise increases and FM threshold is degraded if the IF frequency response is unsymmetric or if the RF signal is not centered in the IF response of the receiver. The effect of unmodulated carrier RFI relatively close in frequency to the desired RF signal frequency is to produce significant noise at the baseband frequency equal to integer multiples of the frequency difference between the RFI and the desired signal. In addition to these beat product noise spikes, overall slot noise is increased at all baseband frequencies and baseband signal suppression starts early. For unmodulated carrier RFI with relatively large frequency offset, the beat products may fall outside the normal baseband frequency range. However, overall slot noise is increased more than for a carrier closer to the desired signal. The effects of the more common degradations of quieting curves are noted on the next page. It should be kept in mind that actual quieting



Typical Areas of Quieting Curve Degradation

Figure 90

curves will move right or left relative to a predicted curve if errors are made in measuring RSL or noise figure. The actual curve will move up or down with errors in estimation or adjustment of receiver per channel rms deviation, consistant error in noise power measurement, or error in determining test point TLP.

9.5 Two common specifications of FM receiver noise performance are 1 dB FM noise threshold and 20 dB noise quieting. Near theoretical FM threshold noise performance can be obtained with hard limiting FM receivers. However, as limiting goes soft, baseband signal and noise threshold performance is affected. For moderate RSLs, as limiting goes soft, additional essentially flat frequency noise due to received carrier amplitude variations is added to the receiver baseband. With moderately soft limiting and moderately weak RSLs, low frequency baseband slot noise is increased causing FM threshold to occur early (for stronger RSL). The slot noise in the high slots is not significantly affected since the noise in those slots is already so great. For very weak RSLs and soft limiting, the noise in all the slots is reduced. This can cause the high slot FM noise threshold to appear better than theoretically possible. When limiting is very soft and RSLs are moderately weak, slot noise is increased in all slots and FM noise threshold occurs early for all slots. For very weak RSLs, however, the slot noise in all slots is reduced. Overall it appears that a baseband slot noise frequency divided by IF bandwidth (f/B) of 0.05 gives the best overall performance in retaining near theoretical FM threshold performance with moderate limiter degradation. As limiting goes soft, FM threshold occurs at increasingly stronger RSLs (relative to theoretical performance) for increasingly lower frequency baseband noise slots.

9.6 The other common noise specification is 20 dB quieting. It seems likely that no receiver achieves hard limiting with no RF input signal. Therefore slot noise noise for very weak RSLs will always be less than the theoretical values. Hence actual 20 dB noise quieting will never approach theoretical values too closely. However, it appears that if the receiver starts to limit as the RSL approaches the 20 dB quieting RSL, the difference in 20 dB quieting RSL for different high frequency baseband noise slots approaches the difference in the theoretical values. For example, the difference in 20 dB quieting RSLs for normalized noise slots of f/B = 0.1 and f/B = 0.05 is 1 dB. As limiting goes soft, the difference in RSL increases. The 20 dB noise quieting RSLs for normalized baseband frequencies f/B less than 0.05 appear to be about the same regardless of degree of limiting. Therefore, the difference in 20 dB quieting for the middle and high slots of a quieting curve appear to be good measures of receiver limiter action.

9.7 The previous measures of FM receiver noise performance related to average noise power (rms voltage level). It appears that the peak to rms voltage level of noise in a 3.1 kHz baseband noise slot is independent of baseband slot frequency or receiver RSL. The peak to rms voltage level is somewhat dependent on peak level definition. However, measurements indicate that the peak to rms level is no greater than about 16 dB.

10. Recommendations

10.1 When a M/W terminal is evaluated, recommend a signal suppression versus received signal level (RSL) curve be obtained for all receivers using the companion M/W transmitter. For standardization a baseband test tone frequency divided by IF bandwidth (f/B) of 0.05 is recommended. Baseband signal suppression is significantly effected by M/W receiver limiting and internal and external RFI. Poor baseband signal performance will reduce path reliability. Also recommend that a set of at least three FM slot noise quieting curves be obtained for each receiver using the companion M/W transmitter. In addition, recommend a spectrum analyzer sweep of each receiver's baseband be obtained with each receiver fully quieted (an RSL of -30 dBm is recommended) by the companion M/W transmitter with radio pilot defeated. Analysis of the fully quieted receiver baseband noise spectrum will identify any spurious noise not previously identified by the quieting curves.

10.2 When LOS M/W terminals are evaluated, it is recommended that three sets of baseband signal suppression and FM slot noise quieting versus RSL be run. The first would be run with a low noise CW M/W RF generator as the quieting source and with the frequency selective voltmeter (FSV) connected to a single receiver baseband output. This would highlight idle receiver problems. The second set would be run with the far end M/W transmitter (with baseband input terminated) as the quieting source and with the FSV connected to the combined receiver baseband output. This would highlight RFI and transmitter induced noise. The last set would again be run with the far end transmitter as the quieting source. However, this time, the transmit baseband would be terminated at the multiplexer end of the baseband cable and receive noise quieting would be measured with the FSV connected to the multiplexer end of the receive baseband cable. A spectrum analyzer baseband sweep with the receiver receiving the normal RSL from the far end transmitter (baseband cable terminated at the multiplex) is recommended to identify spurious signals not identified by the previous quieting curves.

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